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DESIGN CONSIDERATIONS IN THE DEVELOPMENT OF A LOW COST TELEVISION RECEIVER

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DESIGN CONSIDERATIONS IN THE DEVELOPMENT OF A LOW COST TELEVISION RECEIVER

by

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Academic Dean

PREFACE

This paper describes the reasoning used in the development of a low cost television receiver. The work on which this paper is based was done during January, February and March 1948 in the electronics laboratory of the Postgraduate School of the United States Naval Academy at Annapolis, Maryland. The work was done to give the writer experience in practical circuit design problems of the type in which conservation of materials and manpower is of importance.

The writer wishes to express his gratitude to Professor P.E.Cooper of the Postgraduate School for his guidance and assistance in the development and testing of the circuits described in this paper.

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INTRODUCTION

Reception of television programs as broadcast in the United States at the present time requires a receiving device which is considerably more complex than those devices used for reception of most types of information. This complexity is a characteristic of devices which receive information very rapidly and display it in visible form. The receiver is not necessarily more elaborate, nor is the number of parts necessarily larger; but in general the number of different types of circuits involved in a television receiver is larger than the number of different types of circuits involved in other types of receivers.

While the cost of many types of radio receiving apparatus may be continually reduced as the requirements on the apparatus are made less stringent, this is not exactly the case with the television receiver. The sound channel associated with a television receiver might possibly consist of anything from an elaborate multiple diversity receiver on one extreme to a simple crystal detector and earphone combination on the other extreme. The picture channel is quite different. Here even the minimum receiver requires some type of kinescope with its high voltage power supply and control circuits. Some type of sweep generating circuit must be used to provide the raster on which the received information is to be displayed. Some type of circuit must be employed to separate synchronizing information from video information in the received signal. Since the number of circuits employed can not be reduced below a certain minimum the design of a minimum cost television receiver requires the development of a number of circuits each of which can be built at low cost. It is the purpose of this paper to describe the development of some circuits which may be used in a low cost television receiver.

An additional requirement that the complete receiver be easy to build and adjust has been imposed on this development work. This was done in order that the receiver might be constructed by following a written set of instructions in connecting a set of component parts.

Where quality of components is fixed some general idea of the relative costs of two receivers may be gained by simply counting the number of tubes used in their circuits. In general, reducing the number of tubes used reduces the number of other components used, and a reduction in the number of tubes used also makes possible the use of a less extensive power supply system. Reductions of weight and space required generally follow from reduction in number of tubes used. There are, of course, certain cases in which a low cost tube may be used to replace a more expensive transformer as a phase inverter, or a tube diode may prove less expensive than a crystal detector. Several cases of this type will be pointed out.

In starting the design of the receiver the number of functions to be performed will be listed, and a minimum number of tubes will be assigned for each function. Then the design of the circuitry for the performance of the various functions will be investigated in detail.

1	Radio Frequency Oscillator and Converter	1 twin triode
2	Video Intermediate Frequency Amplifier	2 pentodes
3	Video Detector and Amplifier	1 pentode
4	DC Restorer, Sync Separator and Amplifier	1 twin triode
5	Sound Intermediate Frequency Amplifier	1 pentode
6	Sound Detector and Amplifier	1 twin triode
7	Sound Power Amplifier	1 tetrode
8	Complete Sweep Generator	1 twin triode
9	High Voltage Supply	1 diode
10	Low Voltage Supply	1 twin diode
11	Picture Tube	1 kinescope

The above list indicates that all of the necessary functions might possibly be performed by a total of twelve tubes or possibly by eleven tubes if a selenium rectifier were used for the low voltage supply. Such a twelve tube receiver was built and made to perform on signals from local transmitters. A much improved and somewhat less expensive circuit employing a total of fourteen tubes was developed as a final model.

GENERAL CONSIDERATIONS

Before any specific circuits can be discussed it is necessary to make some general decisions as to the nature of the set to be designed. It is easily shown that for lowest possible cost the kinescope to be used should be of the electrostatic deflection type. The maximum brightness possible and the possible definition with bright pictures are not as great in the electrostatic as in the electromagnetic deflection tubes presently available, but the cost of the deflection components necessary for electromagnetic deflection tubes rules out the possibility of using this type of deflection. Of the available electrostatic deflection tubes the 7JP4 is at present the most promising because it is widely used, and for that reason it is low in price. A possible alternate is the 7EP4. If a smaller tube face is adequate the 5BP4 might be used, and if a larger screen is desirable the ten inch SR-1022 produced by Sylvania Electric is a possibility. The design to be presented here is based on the seven inch 7JP4.

The seledtion of the kinescope determines to some extent the bandwidth necessary for the receiver. With a seven inch tube face the largest possible picture with correct aspect ratio of four to three is 5.6 by 4.2 inches. Viewer preference seems to indicate that a slightly larger picture with rounded corners is desirable. A 6.0 by 4.5 inch picture was selected for this receiver. These dimensions do not include the portion of the sweep which is blanked off. With the spot size obtainable on the 7JP4 the average viewer can resolve no more than 240 vertical lines in the six inch picture at a viewing distance of six feet. Thus at a distance of six feet the viewer will be able to perceive no difference at all if the bandwidth of the receiver is made greater than about 3.9 megacycles. When a bright picture is used the spot size of the kinescope increases, and thus the maximum number of lines which can possible be resolved is decreased. This coupled with the fact that average viewer distance is expected to exceed six feet led to reducing the design bandwidth of the receiver to about 2.5 megacycles. It should be noted here that the figure of 2.5 megacycles applies to the final signal delivered to the kinescope. The bandwidth of the radio frequency, intermediate frequency, or video frequency amplifiers need not be 2.5 megacycles individually. It is only necessary that if these units have individual bandwidths which are less than that required that their tuning be so staggered that the product of their individual responses give the proper over all response in both phase and amplitude.

THE LOCAL OSCILLATOR

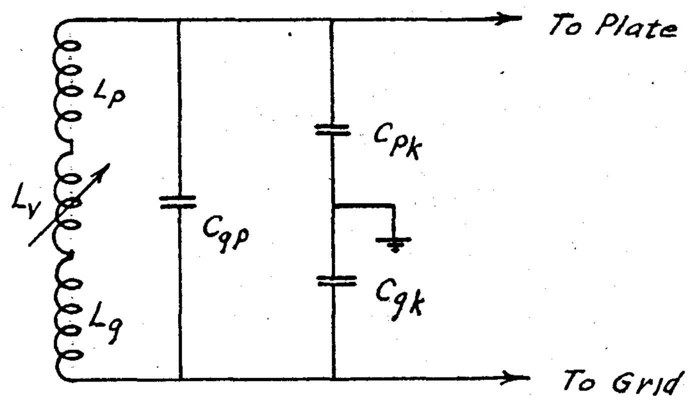
The first actual circuit design to be undertaken is the design of a radio frequency oscillator and converter. Here the problem is as follows: The frequency allocations for the thirteen presently available television channels are:

Channel	Frequency MC.	Sound Carrier	Picture Carrier
1	44-50	49.75	45.25
2	54-60	59.75	55.25
3	60-66	65.75	61.25
4	66-72	71.75	67.25
5	76-82	81.75	77.25
6	82-88	87.75	83.25
7	174-180	179.75	175.25
8	180-186	185.75	181.25
9	186-192	191.75	187.25
10	192-198	197.75	193.25
11	198-204	203.75	199.25
12	204-210	209.75	205.25
13	210-216	215.75	211.25

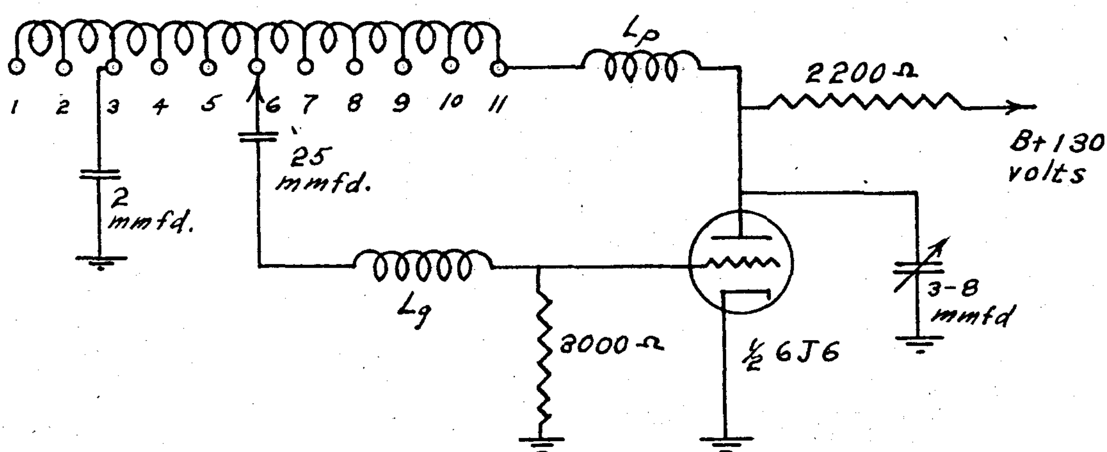
It is desired to change the frequencies of the above channels such that the sound carrier will be centered at 21.25 megacycles and the picture carrier will be centered at 25.75 megacycles. (These are proposed standard intermediate frequencies for television receivers.) Since the oscillator must be higher in frequency than the received signal the oscillator range must be from 71 to 237 megacycles. Several miniature twin triodes which will operate over this frequency range are being produced at the present time. Because of cost and availability factors the two which are to be considered in this work are the 6J6 and the 12AU7. Although the newer 12AU7 is listed at a slightly lower cost than the 6J6, large quantities of 6J6's are available at low cost from government surplus supplies, and this condition has brought about wide use of the 6J6 in new equipment design. A very desirable feature of the 12AU7 from the circuit design standpoint is the fact that this tube has separate cathode connections to the two triodes used in it. If a circuit can be designed such that separate cathode leads are not necessary then the low lead inductance and high transconductance of the 6J6 make it the best of the two tubes from the standpoint of performance.

A number of available circuits employ the 6J6 as an oscillator over the desired frequency range. In general, those employing continuous tuning by means of a variable inductance or capacitor are undesirable in a minimum cost unit because the tuning mechanism does not lend itself to low cost production as well as some other arrangements. The circuit finally selected was the conventional Colpitts arrangement with the stray capacity from plate to ground and from grid to ground furnishing the required center tapped condenser.

Figure one (a) shows the basic oscillatory circuit. Here L_v represents the variable portion of the tank inductance which is changed for the various television channels by changing the position of an eleven position switch. By placing the variable portion of the total inductance in the electrical center of the circuit when the switch is positioned for one of the high frequency channels the effect of the stray capacity of this switch to ground is reduced. When the switch is positioned for one of the six lower frequency channels this stray capacity appears as a part of C_{gk} in the circuit. As shown in figure one (b), a small trimmer condenser from the plate to ground is used for fine turning of the oscillator. This tuning condenser provides more variation than is desirable for the high frequency channels, but reduction in the size of this condenser would reduce the tuning range on the low channels to such an extent that the low channel tuning coils would have to be individually adjusted for the desired frequency. The apparent increase in C_{gk} when the channel switch is turned to one of the lower six channels helps to increase the tuning range on these low channels. It was found that a 2 mmfd. condenser connected from the channel three switch lug to ground gave a further increase in low channel tuning range.



(a) Basic Oscillatory Circuit



(b) Local Oscillator Circuit

Figure 1.

This capacity was chosen to resonate with the total inductance between the channel three and channel seven positions of the switch at a frequency of about 120 megacycles. Thus there can never be a series resonant circuit between the switch slider and ground at the operating frequency of the circuit.

It will be noticed that the switch used is an eleven position device. This was used because such switches are a standard item readily available. With this arrangement it is necessary to tune channels ten and eleven and channels twelve and thirteen on the last two switch positions. Two modifications of the circuits are possible to avoid this. First, if the detent mechanism of the standard type of switch is modified to remove the end stop it becomes a twelve position continuous rotation device. Then, using standard eleven position wafers, another frequency changing coil may be placed between the last of the eleven regular contacts and the lug which is connected to the switch slider. This coil will be connected in the circuit only when the switch is in the new 12th position. Then the coil arrangement may be modified to omit channel one entirely since this channel has not as yet been assigned to any transmitters and since it probably will not be assigned because of interference from other services.

The oscillator uses shunt feed for the plate voltage supply through a 2200 ohm resistor which also serves as a coil form for the inductance L_p. Grid leak bias is developed across a 3000 ohm resistor which serves as a coil form for L_g. Direct Current isolation between grid and plate circuits is provided by a 25 mmfd. coupling condenser. The particular arrangement of this condenser between the switch contact and L_g was chosen because in this connection the wire lead of the coupling condenser could be wound around the grid leak resistor to form L_g. The scheme of using these resistors as coil forms was found desirable because it helped to minimize and keep constant the lead inductance which became very important in the high frequency range of the oscillator. The values for the resistors used were arrived at by trial. The grid leak resistor size must be low enough to prevent blocking. With the 6J6 tube an RC product of well under one micro-second is desirable for the grid circuit, and the minimum size of the coupling condenser is determined by the requirement that its reactance be low at the lowest oscillator frequency. The resistance values shown in figure one (b) prevent oscillator blocking and help to stabalize the output voltage of the oscillator over its tuning range.

THE MIXER

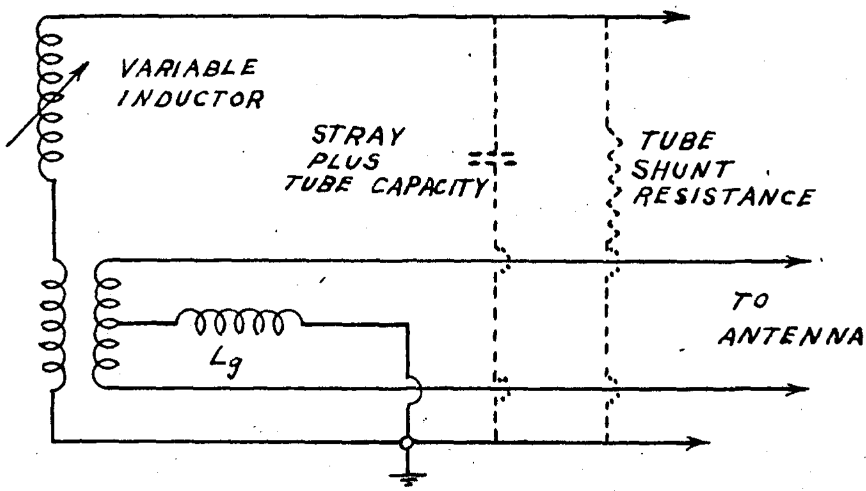
The remainder of the frequency conversion circuit is shown in figure two. It is basically a grid leak detector with both local oscillator and signal voltages injected in the grid circuit. The oscillator voltage is injected by capacity coupling to the plate of the oscillator tube. The main problem met in the design of this circuit is that of finding some reasonably efficient method of coupling the signal from the antenna to the mixer grid. To do this it is necessary to find first the equivalent input admittance of the mixer grid. Since the output circuit of the tube is tuned to an intermediate frequency much lower than the received frequency the output impedance which the tube sees at signal frequency can be written:

$$Z_{out} = \frac{1}{j\omega C_{out}}$$

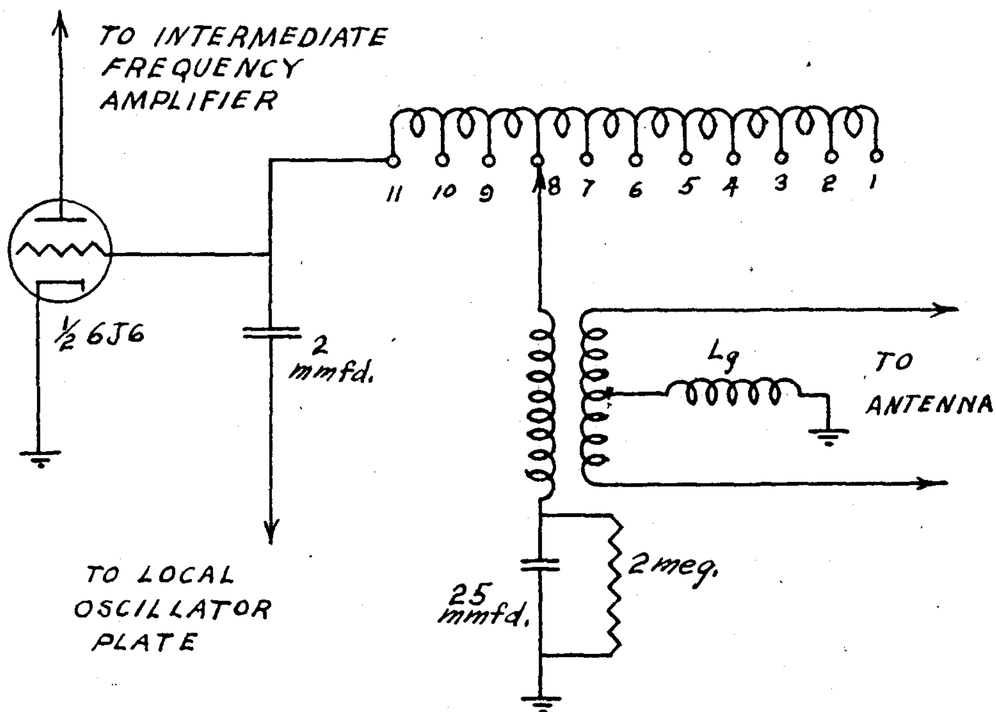
The input impedance of the tube due to the presence of inductance in the cathode lead and grid-cathode capacity is given by:

$$Z_{in} = \frac{j\omega L_K\left(\frac{\mu}{j\omega C_{gK}} + Z_{out} + r_p\right) + \frac{1}{j\omega C_{gK}}\left(r_p + Z_{out} + j\omega L_K\right)}{r_p + Z_{out} + j\omega L_K}$$

Assuming a value of 5×10^{-8} henrys for L_k and substituting appropriate values for the other quantities shows that there is a reactance term equivalent to the shunt



Input Coupling Circuit



Frequency Converter

Figure 2.

reactance presented by the grid cathode capacity. Due to the cathode lead inductance a shunt resistance appears across the input. Its value varies inversely with the square of frequency from about 6000 ohms in channel two to about 1500 ohms in channel thirteen.

If Cathode lead inductance is neglected and the presence of grid-plate capacity is considered the input admittance is given by:

$$Y_{in} = j\omega C_{gp} \frac{j\omega C_{out} + \frac{\mu+1}{r_p}}{j\omega(C_{out} + C_{gp}) + \frac{1}{r_p}}$$

The loading due to grid plate capacity, if computed, is found to be nearly independent of frequency; and it is less than that due to cathode lead inductance. A complete treatment of the input admittance of the tube involving all of the internal capacities of the tube and the various lead inductances of the tube is of doubtful value because of the great uncertainty which exists as to the exact distribution of the various capacities and inductances.

There are other positive conductance terms in the input admittance due to electron transit time effects. The transit time effect is practically negligible in the low frequency channels, but it tends to increase the effective value of the shunt conductance in the high frequency channels. For purposes of further calculations an input circuit consisting of a condenser of eight micro-

microfarad capacity in parallel with a resistor will be assumed. The value of the resistor will be assumed to vary inversely with the square of frequency from 1000 ohms in channel thirteen to 6000 ohms in channel two.

With no extra loading the band width of this circuit, which is given by BW = 1/6.28 RC, is twenty megacycles at channel thirteen and four megacycles at channel two. From these figures on band width it can be seen that there should be little loss in performance due to the use of only two switch positions for the four highest frequency channels.

The basic coupling circuit to be investigated is shown in figure three. Here the condition for maximum energy coupling from the antenna to the grid circuit is given by:

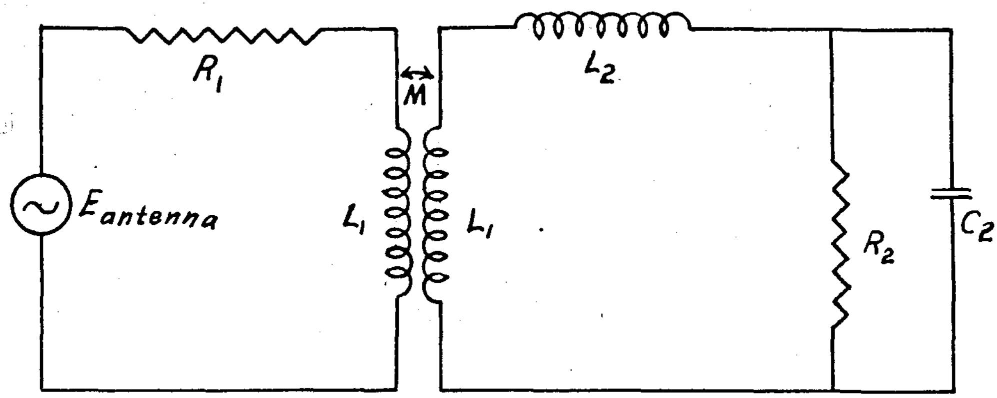
$$R_1 - j\omega L_1 = \frac{(\omega M)^2}{Z_{sec}}$$

Where R_2 is greater than $5X_{c2}$ The resistive component of Z_s is approximately

$$R_s = \frac{1}{\omega^2 C_2^2 R_2}$$

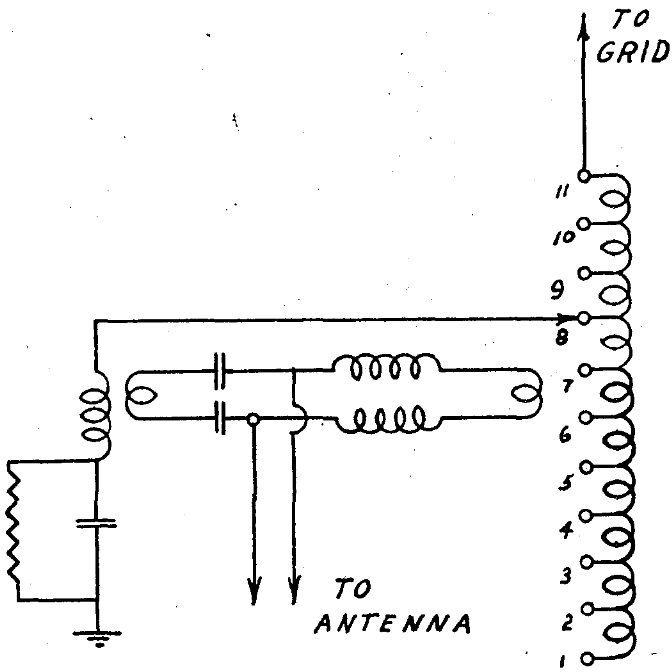
It can be shown in this case that if the band width is such that

$$R_1 \omega^2 C_2 \gg \omega_{BW}$$



Equivalent Antenna Coupling Circuit

Figure 3.



Coupler With Improved Low Frequency Response

Figure 4

then the secondary circuit must meet the usual resonance equation

$$\omega^2 = \frac{1}{(L_1 + L_2) C_2}$$

for maximum energy transfer. Since this is equivalent to saying that only the resistances shall be matched for best results, the equation which should be satisfied is

$$R_1 = (\omega M)^2 \omega^2 C_2^2 R_2$$

Solution of this equation gives a value of M of 0.2 microhenrys for channel two and 0.038 microhenrys for channel thirteen. A further restriction is placed on M in that L_1 must be a large portion of the secondary resonant circuit on high channels. If L_1 can be made eight tenths of the total secondary inductance and if the two coupled coils marked L_1 in figure three are interwound closely the value of 0.038 microhenrys for the mutual inductance can be obtained. With this arrangement the coupling circuit gives optimum results over all of the high frequency channels. On the low frequency channels the coupling system gives from about 15 to 50 per cent of optimum performance as frequency is changed from channel two to channel six.

Several other arrangements for coupling the antenna signal to the grid of the mixer have been examined. The idea of using eleven different coils each with its own coupling coil was abandoned because two more switch wafers would be required to switch coupling coils. This and the large number of coils would complicate the mechanical construction problem greatly. One simple change which can be made to the circuit shown in figure two to improve its performance in the low bands is the addition of some type of coupling between the antenna circuit and the jump coil between switch positions six and seven. One possibility is shown in figure four.

For minimum cost and least circuit complexity the circuit shown in figure two was used in receivers which were built in this investigation. One modification was made to this circuit to obtain improved low channel performance. This consisted of making the antenna coil of the coupling circuit slightly larger than the grid coil of this circuit. Bandwidth measurements on completed receivers indicate that the grid loading in the low frequency channels is not so great as the approximate figure given above. This results in greater low channel coupling efficiency and less desirable bandwidth characteristics.

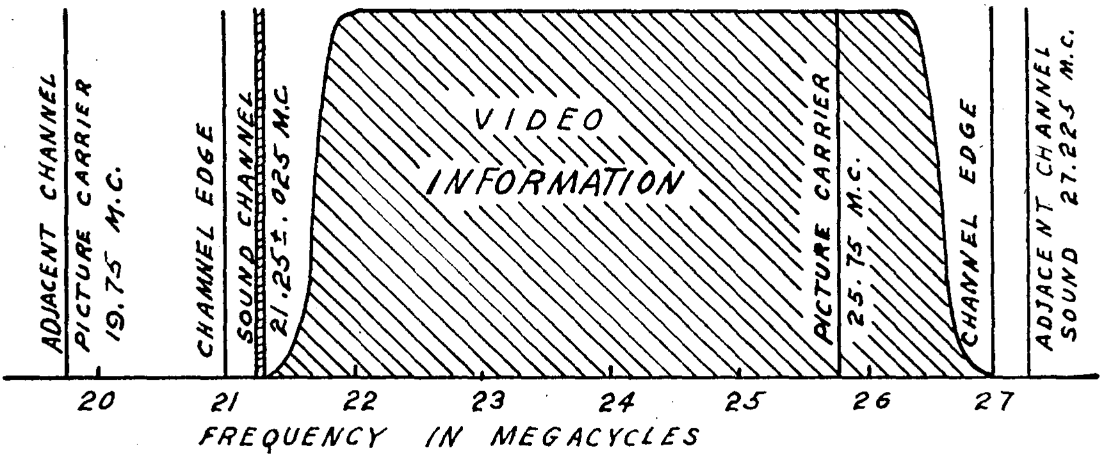
It should be noted here that if the coupling is to be poor in either the high or low channels it should be made best in the high channels. This is due to the fact that for a signal to be received with an antenna with a fixed number of elements (usually from one to three) the antenna output voltage will be five to ten times greater for the low channels than for the high channels if the field strength is fixed. This is due to the fact that the antenna cut for the low channel is much larger than one cut for the high channels.

Besides coupling efficiency at the signal frequency there is one more requirement for the coupling system. That is that it should provide low coupling efficiency at the intermediate frequency of the receiver, and it should provide low coupling efficiency for all in-phase signals applied to the antenna terminals. It is assumed in the development of this tuner that the antenna will be connected to the tuner by some sort of balanced line such as the low cost 300 ohm twin lead now available. A balanced twin line such as this has the disadvantage that the two conductors act together as a long wire antenna and couple undesired in-phase noise and intermediate frequency signals to the two antenna posts. One particularly objectionable form of noise is that produced by automobile ignition systems as impulse noise of this sort acts as a false synchronizing signal to tear entire blocks of scanning lines out of the picture. Very strong intermediate frequency interference is produced by both foreign and domestic short wave transmitters using frequencies between twenty and twenty-seven megacycles. Both of these types of interference are most severe in the low channels since the noise voltages produced at these frequencies may be quite large and since the grid circuit of the mixer tube has appreciable reactance at the intermediate frequency when it is tuned to a low channel.

The coupling scheme shown in figure two has rather low efficiency in the lowest frequency television channel, and its coupling efficiency at the intermediate frequency is quite low. It was found that by connecting the center of the antenna coupling coil to ground a very great improvement in rejection of in-phase signals at all frequencies could be obtained. One disadvantage to the grounded center tap is that it greatly increases the effective value of the stray capacity across the tuning coil on the high channels. A compromise can be made here by making the grounding lead long enough so that it has an inductive reactance greater than the reactance of the capacity between the high end of the grid coil and the antenna coupling coil. This is shown as inductance L_g in figure two.

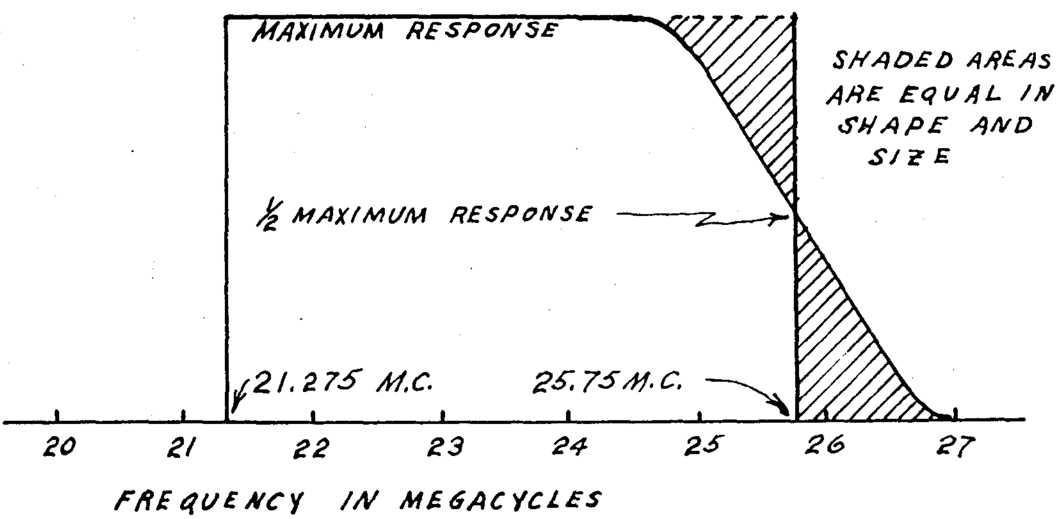
THE VIDEO INTERMEDIATE FREQUENCY AMPLIFIER

The second element of circuitry to be developed is the video intermediate frequency amplifier. Using the frequencies 21.25 megacycles for zero deviation sound carrier and 25.75 megacycles for video carrier the signals presented to the video intermediate frequency amplifier are as shown in figure five. It can be noted that the transmission system is of the vestigal side band type. In terms of the intermediate frequency the upper side frequencies have been attenuated greatly such that the side frequencies above the channel edge at 27.0 megacycles are missing. Such a signal can be detected without distortion by ordinary means provided a network which has the characteristics shown in figure six is inserted ahead of the detector. This response can be thought of as reducing the amplitude of the carrier by one half with respect to the amplitude of the lower side frequencies. Thus when only the lower side requencies are mixed with the carrier in detection the original percentage of modulation is preserved. In reducing the carrier amplitude some of the low side band frequencies near the carrier are reduced in amplitude. To compensate for this the response curve is made to include high side band frequencies as shown in the shaded areas in figure six. An important fact to be observed here is that with the picture carrier tuned to the half way point on the response curve as shown in figure six the output video voltage will not be the maximum possible and the picture received will not be the brightest possible. This can be explained by the fact that if the local oscillator of the receiver is detuned such that the picture carrier comes in at say 24 megacycles the detector will see a conventional double side band signal near the carrier. Since a large part of the energy of the transmitted signal is within about one megacycle of the carrier this double side band detection will nearly double the output video voltage. The side frequencies removed by more than one megacycle from the carrier will now be greatly distorted in detection, for there are no upper side band frequencies present in this range since they were attenuated at the transmitter, and the carrier is now twice as large as needed for detection of the lower side frequencies alone. In addition to this a large portion of the lower side band as transmitted has now been moved outside the pass band of the video intermediate frequency amplifier. Since it is this lost portion of the lower side band which would have produced the high frequency components of the detected video signal the picture will suffer loss of definition. Another difficulty met in this type of tuning is that the frequency of the sound carrier



Intermediate Frequency Spectrum

Figure 5



Desired Intermediate Frequency Response

Figure 6.

has now been moved lower and outside the sound channel of the receiver. The television viewer notices the above condition in that when he tunes his receiver to a weak signal he does not get picture and sound on the same setting of the tuning control. In localities where all received television signals are low in amplitude it is possible to tune the receiver for acceptable brightness of the picture and retune the sound channel of the receiver to make its pass band match the sound frequency. This is what is usally done in construction of television receiver kits which are tuned to provide reception on minimum signals. The loss of definition in the picture is then tolerated because this type of tuning simply makes the difference between useable pictures and no pictures at all.

In figure six the response is shown as dropping sharply at 21.275 megacycles which is the high frequency edge of the sound channel where a deviation of plus or minus 25 kilocycles is used. Where the intermediate frequency bandwidth is to be decreased in order to get increased gain from a given number of tubes this figure of 21.275 megacycles would be increased. The sharp drop, at whatever the lower cut off frequency is, is desirable only when the phase response of the band pass network can be controlled in such a manner that no distortion is introduced in the transient response of the amplifier.

In general this can not be done in a simple video intermediate frequency amplifier.

This statement can be partially explained by the following line of reasoning: Suppose that the band pass characteristics of the video intermediate frequency amplifier are to be obtained by cascading a number of circuits having simple parallel resonant circuit responses. Then the over all response is the product of the individual responses. Further, let there be only one response curve considered for each resonant frequency; that is, let the resonant frequencies of all the cascaded networks be staggered through the pass band. (If more than one network is tuned to one frequency consider the composite for these networks as the response for this particular frequency.) Now, the only combination of response curves which will produce the abrupt drop and sharp skirt at 21.275 megacycles as shown in figure six is one in which one or more of the individual networks has a very sharp resonance curve at or slightly above 21.275 megacycles. The presence of any network having a sharp resonance in the cascade chain is the same whether this sharp resonance is produced by feedback around an amplifier tube or by low energy dissipation in the passive elements of which the network is composed. Consider, for example, an intermediate frequency amplifier which has in its response a sharp peak at 24.75 megacycles. Let the picture carrier be correctly tuned to 25.75 megacycles. When picture elements are transmitted which require a sudden change from light to dark all resonant circuits in the intermediate frequency amplifier are shock excited. If all of these circuits have Q's which are of the same order of magnitude the output of the amplifier will faithfully reproduce its input. The high Q circuit which produces the peak at 24.75 megacycles will continue to ring for a large number of cycles due to the shock excitation alone. In the detector circuit the 24.75 megacycle signal will be mixed with the 25.75 megacycle carrier to produce an extraneous video signal at a frequency of one megacycle which may last for several cycles. Since one horizontal trace represents 63.5 microseconds, a one megacycle video signal will produce vertical lines which are 0.11 inches apart on a six inch picture. The effect of a sharp peak on the response curve at 24.75 megacycles would then be to reproduce a series of one or more low definition positive or negative images of high contrast vertical elements of the transmitted picture. The effect will, in general, be more noticeable on changes to white from black. This is due to the fact that the large transmitted signal for black picture elements may mask the extraneous signal during black elements. A sharp respnance curce at 21.275 megacycles as required for the response curve of figure six would produce an extraneous ringing signal at a frequency of about 4.5 megacycles. The effect of this would be most easily noticed as a blurring of the center portion of the vertical wedge of lines on a standard test pattern.

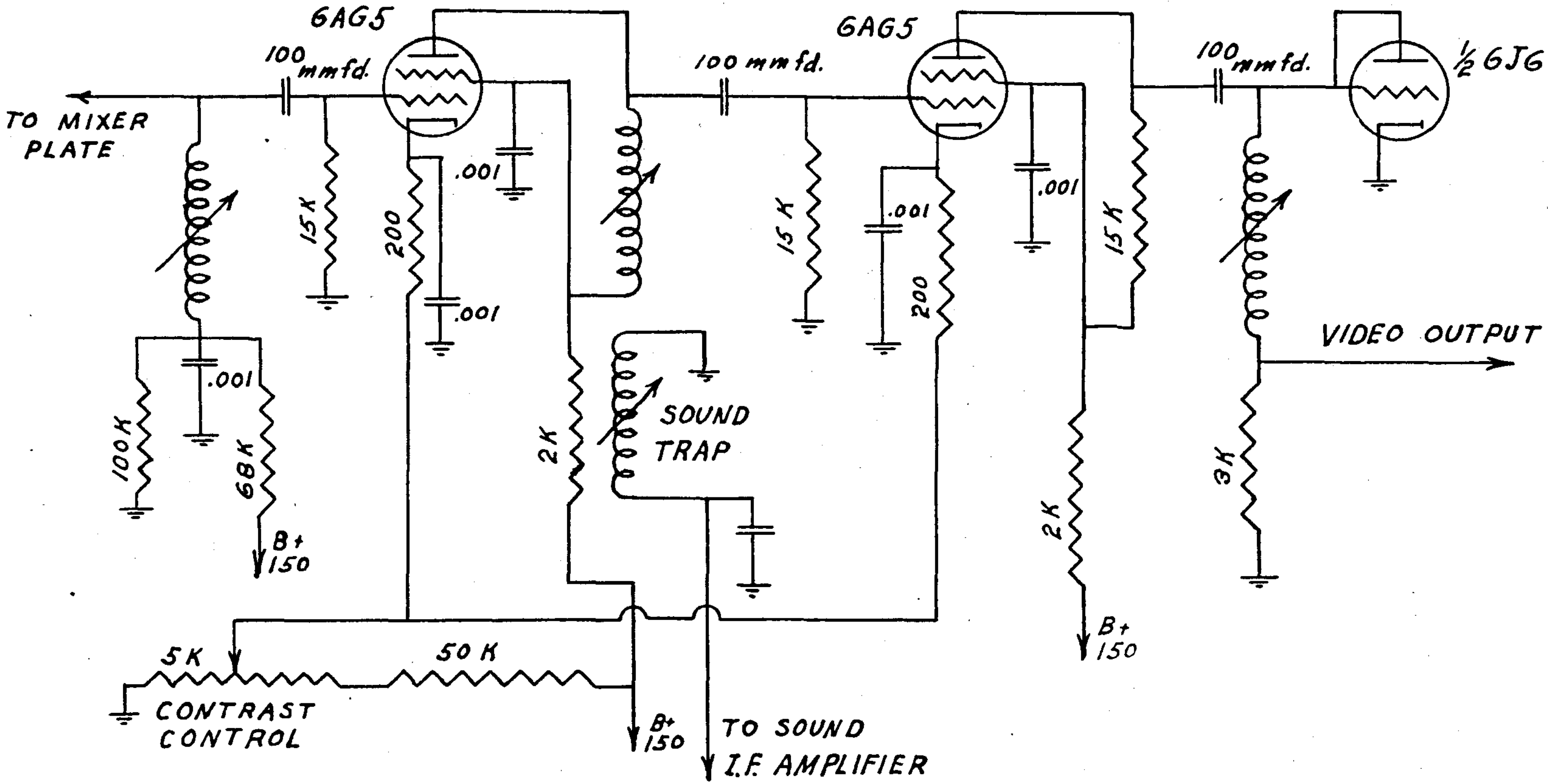
A more satisfactory explanation of the above may be had by writing the product of the response characteristics of the cascaded networks involved in a video intermediate frequency amplifier and examining the phase response in the region of any sharp discontinuity in the amplitude response curve. The discussion above is given chiefly because it shows an easy method of recognizing regenerative peaks in the intermediate frequency amplifier response by their effect on the picture.

Since only two tubes have been assigned for use in the video intermediate frequency amplifier it might seem logical to use some type of transformer coupling since this type of coupling leads to the greatest possible gain bandwidth product. However, the stagger tuned intermediate frequency amplifier has several features which seem to make it the most desirable. The stagger tuned coupling system is the least expensive possible. It has adequate flexibility to allow the intermediate frequency response to be distorted to correct for tuner and video amplifier response. By making the first stage of the video intermidiate frequency amplifier the low frequency member of the doublet it can be used to give gain at the sound channel, and the sound can be separated from the video signal between stages.

In this particular case the video intermediate amplifier can not be easily designed by use of the usual formulas for stagger tuned amplifiers. This is due to the fact that the desired response curve is not the square top symetrical curve usually discussed in the literature, and it is also due to the fact that one of the tuned circuits has a load resistor which is determined by the video detector. The final intermediate frequency amplifier design was developed by building the set and trying various loadings and tuning inductances. The video intermediate frequency amplifier and video detector are shown in figure seven. The contrast of the picture is controlled by varying the gain of this amplifier. Since the sound trap which separates the sound channel from the video is placed between the two amplifier stages variation in contrast of the picture will effect the sound volume. This is not desirable, but the alternative of controling the contrast by changing the bias on only the second amplifier tube leads to excessive changes in band pass characteristics with variation of contrast control setting.

The tuning inductors used in this circuit are of the variable iron core type, and, although it increased their cost, they were individually shielded in aluminum cans. This was done to minimize the possibility of regenerative peaks in the overall response. Shielding of these coils also makes the set much easier to align when constructed in kit form.

Of the three tuned circuits used the one in the detector input is of lowest Q. This is necessary since the RC product in the video detector must be kept small, and the input capacity of the video amplifier is of the order of twenty micromicrofarads. Because it is of low Q the detector input circuit was used to give the sloping response required at the high frequency end of the response curve. The first tuned circuit was tuned to the low frequency end of the response curve so it would not too greatly attenuate the sound frequencies. The one remaining circuit was tuned to give the desired response in the middle of the curve. If all three circuits are tuned to nearly the same frequency the amplifier will oscillate. This is undesirable in a receiver which is to be constructed from a kit of parts. The difficulty may be partially overcome by making the tuning range of the coils such that the detector will tune only from mid band upward and the first amplifier grid coil will tune only from mid band downward in frequency.



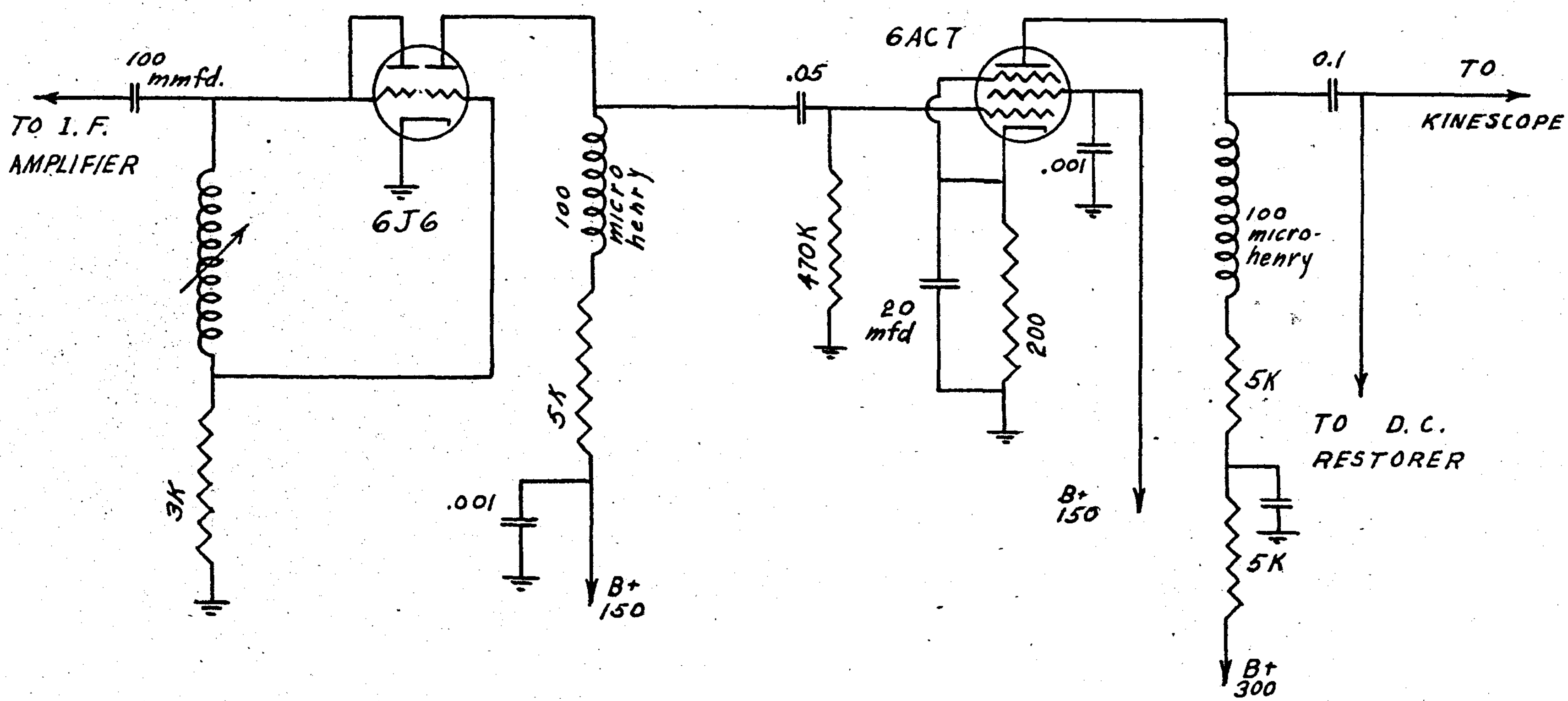
Video Intermediate Frequency Amplifier & Detector
Figure 7

THE VIDEO DETECTOR AND AMPLIFIER

The third element of circuitry developed in this investigation was the video detector and video amplifier. A number of different possibilities are to be considered here. In view of the assignment of only ohly one tube for this job the natural choice of circuits leads to a video amplifier which will act as a grid leak detector also. Such a device will work, but in order to pass the higher video frequencies some loss must be accepted in detection efficiency. A more basic difficulty is met in the nature of the output voltage of a grid leak detector. In television signals as broadcast at present maximum signal intensity corresponds to black on the kinescope. In a grid leak detector maximum signal input corresponds to peak positive output. If a positive signal is to give darkened trace on a kinescope it must be applied to the kinescope cathode with the kinescope grid held at constant potential. This sort of connection is possible, even though not desirable, with some cathode ray tubes. Unfortunately, some tubes, the use of which may be desirable, have their cathode and heater connected internally. (Examples are the 7EP4, the 5BP4, and the 5BP1-A. The last tube listed is a green screen type, but it is very popular because it is available in large quantities at very low cost on the surplus market.) Even with a separate transformer winding to heat the kinescope filament the capacitive loading presented by the cathode of such a tube is prohibitively low.

If a grid leak detector is not to be used because its output is of the wrong polarity the next logical move is to consider the use of something such as the popular germanium diode type 1N34 as a detector followed by an amplifier with its output connected in a manner such that the video output is negative for peak signal. This output can then be connected to the kinescope grid in the usual manner. Improved detection efficiency and increased video gain are possible with this arrangement. From cost considerations and even more attractive proposition is to use a twin triode such as the 6SN7-GT as a diode or grid leak detector and use the second section of this tube to perform the functions of DC restoration and separation of the synchronizing signal. This scheme was tried in a model of the receiver. As might be expected, some trouble was experienced with coupling between the two sections of the tube. Since the DC restorer portion of the tube functioned only during the synchronizing signal it was found possible to keep the circuit stable so long as the gain of the grid leak detector was kept low.

In building the final model of the receiver it was decided to accept the slight increase in cost brought about by an extra video amplifier stage and the circuit shown in figure eight was used. It is interesting to note that there is very little difference between the cost of this circuit and the cost of a 1N34 detector followed by a 6AC7 video amplifier if the 6J6 tube is purchased from the large surplus stocks now available.



Video Detector And Video Amplifier

Figure 8.

THE D. C. RESTORER

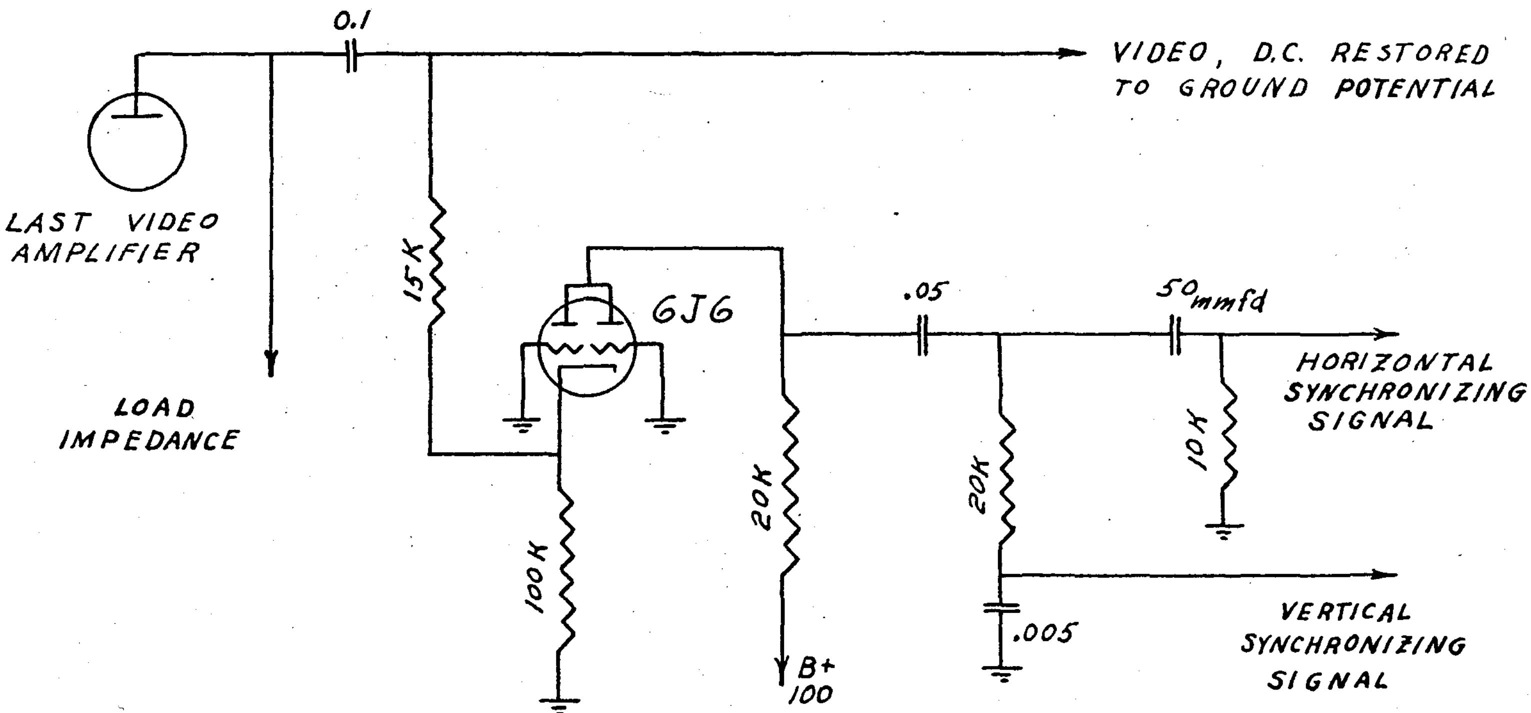
The output of the video amplifier shown in figure eight is an alternating voltage which varies about some average value. For a dark picture its peak to peak voltage is low, and for a bright picture its peak to peak voltage is high. Assume that a bright picture is being received and that the intensity control on the kinescope has been set for proper brilliance. That is, the average value of the grid cathode voltage of the kinescope has been adjusted. Now let the transmitted picture be changed from bright to dim. Since the kinescope average voltage has been set the average brilliance of the kinescope will remain fixed and the reduced output voltage of the video amplifier will cause less variation about that average. It can be seen that if average picture brilliance is to correspond to average illumination of a scene a direct current component must be injected into the output of the video amplifier.

At the television transmitter the modulating voltage is adjusted such that the completely black scene corresponds to 75 percent of the maximum power output of the transmitter. Maximum power output of the transmitter is used during the peak of the synchronizing pulse. For any light the carrier amplitude is modulated downward from the 75 percent power point, and maximum white picture is made to correspond to about 15 percent of maximum power. If in the receiver the peaks of the synchronizing pulses are made to bear a fixed relation to ground potential such that the output voltage of the video amplifier can move but one way from this fixed reference voltage then this reference voltage can be set such that a signal of 75 percent peak power corresponds to completely black picture. The average illumination of the picture on the kinescope will then correspond to the average illumination of the picture being transmitted.

There are two general systems of clamping the peaks of the synchronizing pulses to a fixed reference. One very attractive system is to use grid leak bias on the final video amplifier and connect the grid of the kinescope directly to the plate of this amplifier. The cathode of the kinescope is then moved positive with respect to ground far enough to secure the desired brilliance. Since the grid of the final amplifier is clamped to ground potential by the diode action of the grid and cathode of the tube the output voltage of the amplifier is clamped to a fixed voltage. This scheme falls down in one important respect. Suppose that it is desirable to use a load resistor of 5000 ohms with a low frequency boost circuit employing a 5000 ohm resistor and a 10 microfarad condenser in the B voltage supply. Reference to the characteristic curves for a 6AC7 show that the power supply voltage must be something over 400 volts if clipping of negative output peaks is to be avoided. For economy in the power supply it is desirable to keep the required power supply voltage to below 300 volts.

The second possible scheme of DC restoration requires no compromise in the design of the video amplifier. It consists of connecting the output lead of the video amplifier to the cathode of a diode the anode of which is connected to a fixed reference voltage. A direct current path to ground from the cathode of this diode is provided through a high resistance. In order to reduce the capacitive loading of the diode cathode on the output of the video amplifier the connection here can be made through a resistor of a few thousand ohms. If the clamping diode here is made of the grid and cathode of a triode tube and the plate of this triode is returned through a load impedance to a source of low positive voltage, a negative pulse will appear at this plate each time a synchronizing pulse causes the diode to conduct. Thus the DC restorer can be made to separate the synchronizing pulses from the video signal.

The diagram of the complete direct current restorer, synchronizing separator, and synchronizing amplifier is shown in figure nine. In this circuit the composite synchronizing signal is passed to a differentiating and an integrating circuit. The output of the differentiator



Direct Current Restorer & Synchronizing Signal Separator

Figure 9

is the horizontal synchronizing signal, and the output of the integrator is the vertical synchronizing signal. A compromise is necessary in the plate voltage for the separator tube. For great output voltage this voltage should be high, but with high plate voltage conduction will start in the tube on the edge of the blanking pulse of a weak signal. This is undesirable because in this condition the trace will jump out of synchronization when a very dark picture element is present near the right hand edge of the picture. The best compromise value is one in which the amplitude of the synchronizing pulse is large enough to cut the tube off when the video output is great enough to reproduce the minimum acceptable picture.

In the circuit shown in figure nine the negative peaks of the video signal are clamped to ground potential. Kinescope brilliance can be controlled in this case by returning the kinescope cathode to a potentiometer which supplies a voltage positive with respect to ground. If it is desirable to return the kinescope cathode to ground the grid and cathode ground return of the D.C restorer can be returned to a variable negative voltage for brilliance control. If a 5BP4 or a 7EP4 type tube is to be used it is necessary, because of the internal connection between cathode and filament, to use this last scheme if the filament of the kinescope is to be heated from the common heater supply. If a negative voltage for the kinescope bias is not already available the circuit of figure nine can be used and the filament of the kinescope can be supplied from a separate winding on the power transformer.

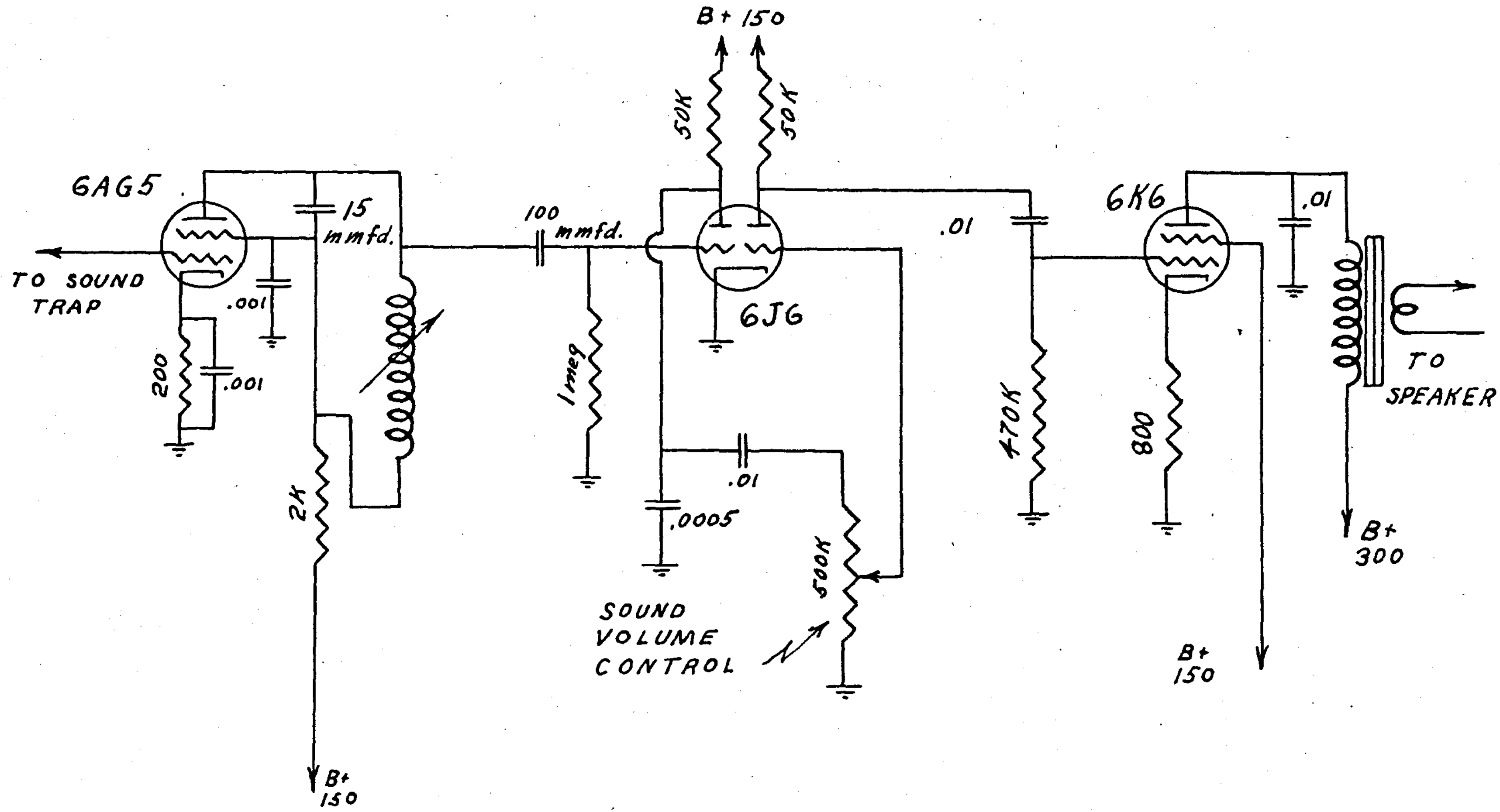
With the addition of the circuit shown in figure nine the video signal chain of the receiver is complete. Provisions have been made for changing the frequency of the incoming signal, amplification and restriction of its bandwidth at an intermediate frequency, detection and amplification of the video information of the signal, restoration of its direct current component, and separation of its synchronizing information. The sound channel which accompanies the picture will now be considered.

THE SOUND CHANNEL

Before any specific design work can be done on the sound channel it is necessary to consider some basic factors. The sound information is broadcast by frequency modulation of a carrier which is entirely separate from the picture carrier. In most cases the power used for the sound carrier is equal to the peak power used for the picture transmitter. The sound channel can be handled as a conventional frequency modulation system except that it is common practice to use much less deviation in Television work than in frequency modulation broadcasting. This restriction is placed on the deviation to prevent sound interference from appearing on the picture of high definition receivers without requiring an extensive sound trap in the video chain.

A large number of methods are available for detecting the information in a frequency modulated signal. The ratio detector would seem to be one logical choice here since it requires no limiting prior to the detector. The output voltage of a ratio detector for a given input voltage is rather low. The more conventional discriminator circuits might furnish more output but less quieting than the ratio detector. Both of these types of detector have the disadvantage of being difficult to align properly unless some test equipment is available. A great improvement in sensitivity and ease of adjustment can be achieved by use of a high gain slope detector instead of one of the discriminators. Some reduction in cost can be realized in a slope detector because a single high Q coil can be used instead of a more expensive discriminator transformer. A slope detector was used in the receivers built in this investigation.

Figure ten shows the complete sound channel for one of the receivers studied. The intermediate frequency signal from a sound trap (see figure seven) on the video intermediate frequency amplifier is amplified by a 6AG5 tube as sound intermediate frequency amplifier. If the receiver is to be used near a television transmitter where large amplitude signals are available it would be desirable to use low plate voltage on this stage in order to get some limiting action. The plate circuit of the intermediate frequency amplifier is tuned to a frequency below the zero deviation frequency of the sound carrier such that this zero deviation point is on a linear portion of the resonance curve. The resonance point is placed below the carrier in order to keep the peak of the audio response away from the amplitude modulated video signal. Because of the low deviation used in television the tuned plate circuit may have a Q of the order of 100. The sound trap in the input of the sound intermediate frequency amplifier may have a similar value of Q.



Sound Channel With Slope Detector

Figure 10

(Note that if this sound channel is to be used for reception of standard frequency modulation broadcasts employing up to 150 kilocycles peak to peak deviation these tuned circuits must have lower Q, and the signal output on television sound will be reduced by about 60 percent.)

The tuning procedure for the slope detector is very simple. The receiver is first aligned to receive best picture. Then the tuning slug in the plate coil of the sound intermediate frequency amplifier is moved back and forth until a position for best sound quality is found. The tuning slug for the sound trap is then moved to a position for best volume. These adjustments may be repeated until best over all sound is obtained.

The slope detector is followed by a high gain amplifier using the second half of the 6J6 detector tube. This drives the final power amplifier which is a 6K6. Reduction of the screen voltage of the 6K6 reduces its plate current requirements without decreasing its output too much. This power amplifier drives a small permanent magnet type dynamic speaker. The low frequency characteristics of this speaker are poor, and the sound signal as broadcast has its high frequency components pre-emphasized by a 100 microsecond RC network as in standard frequency modulation broadcasting. The effects of both of these factors may be compensated by a small condenser shunting the output transformer of the power amplifier.

THE SWEEP GENERATOR

A major portion of the laboratory work done in the design of the low cost television receiver described here went into the design of the next major portion of circuitry, the sweep generator.

Two sweep voltages are required for deflection of the spot in the electrostatic kinescope. One deflecting voltage, has a basic frequency of 15750 cycles per second; and the other, the vertical sweep voltage, has a basic frequency of 60 cycles per second. Both voltages are of saw tooth wave form, and in both cases the flyback time of the saw tooth should be less than approximately five percent of the period so that the retrace function in the kinescope will be performed during the transmitted blanking pulse. Since the sensitivities of the two sets of deflecting plates of a kinescope differ approximately in the ratio 4/3, and since this is the desired aspect ratio for the picture, it is possible to connect the kinescope so that the magnitudes of the two deflecting voltages can be made approximately equal. The required magnitude of the sweep voltage will differ greatly with different tubes and different accelerating voltages. For the 7JP4 tube used in this work an average value of thirty volts per inch of trace per kilovolt of accelerating potential may be used for the horizontal trace. The trace length may be considered six inches.

One further requirement on the sweep voltages is that they should be individually balanced with respect to ground. This is necessary if fine focus is to be obtained throughout the length of the trace. It is not necessary that the voltages be exactly balanced, however, as a small amount of unbalance can be made to result in only a small amount of defocusing on the edges of the picture where it is not particularly objectionable.

The conventional circuit used for generating the sweep voltage is a condenser resistor combination with a tube used as a switch to charge or discharge the condenser. The exponential change of voltage across one of the passive circuit elements is essentially linear over a small range, and this voltage is used as a saw tooth sweep. If the B voltage supply is used to generate the sweep voltage the maximum peak to peak voltage change for a linear saw tooth wave is about 70 volts or about 1/5 of the B supply voltage. Several types of switching circuits are available for the sweep generator. A blocking oscillator may be considered as a switch which is momentarily closed during its cycle of operation. One side of a cathode coupled multivibrator will perform the same switching function without requiring a blocking oscillator transformer.

Regardless of the type of sweep generating circuit used the sweep voltage must be amplified after it is generated. Most of the sweep generators in common use generate a saw tooth voltage which is unbalanced with respect to ground. The amplifier following the sweep generator is then made to act as a phase inverter to produce an approximately balanced voltage which can be applied to the kinescope deflection plates. The maximum output voltage of such an amplifier is approximately 75 percent of twice the B supply voltage. For a three hundred volt B supply a sweep voltage of about 450 volts may be produced. In the case where this is used for deflection of a 7JP4 kinescope the accelerating voltage of the kinescope is limited to approximately 2500 volts. This is undesirable since at 2500 volts accelerating potential the maximum possible focusing ability and contrast range can not be realized in a 7JP4 tube.

Where standard circuitry is used in a multivibrator type raster generator eight triode tubes are required to produce two sweep voltages in final form. The maximum amplitude of the sweep produced is limited to something below the desired value. Some distortion is introduced in the sweep voltage due to the use of an amplifier working at nearly maximum possible output voltage.

The possibility of using a transformer to provide increased deflection voltage must not be overlooked. If a transformer is used the wave form from the sweep generator must be a trapezoid to take care of the leakage inductance of the transformer. Linearity controls must be provided to adjust the shape of this trapezoid when it is desired to adjust raster size. The vertical transformer must have a heavy iron core to prevent distortion of the low frequency sweep. A high quality powdered iron core with a litz wire winding should be used for the horizontal sweep transformer since it is necessary to pass up to the tenth harmonic of the 15750 cycle per second sweep voltage in this circuit. A diode dampling circuit should be used with this transformer in order to secure fast retrace. The use of transformers makes possible higher deflection voltages, but it is not an inexpensive answer to the problem.

It should be noted here that the problem of producing large sweep voltages is even greater if the television receiver is to be made of the a.c.-d.c. type with B voltage supply limited to 115 volts when operating on direct current.

Several of the difficulties met in the production of the raster can be overcome if a very high voltage power supply is available and if the sweep voltages can be generated at the correct magnitude and applied directly to the kinescope deflection plates without being amplified. Such a high voltage supply is always available as the accelerating voltage supply. Since the maximum deflecting voltage required is seldom greater than 20 percent of the accelerating potential used it should always be possible to generate a saw tooth voltage of sufficient magnitude for direct application to the kinescope. It is only required then that some circuit be devised which will generate such a voltage in a balanced form.

The basic circuit to be considered here is shown in figure eleven. In this circuit consider first the case where switch s is closed and has been closed for a long time such that

$$(1) \qquad V_{AC} = V_{BC} = \frac{R_2}{R_1 + R_2} E$$

gives the voltage of the two points A and B with respect to C. Now let the switch be opened at time t=0. C_1 discharges through R_1, and C_2 discharges through R_2. During this period the potentials of points A and B with respect to C are given by the equations:

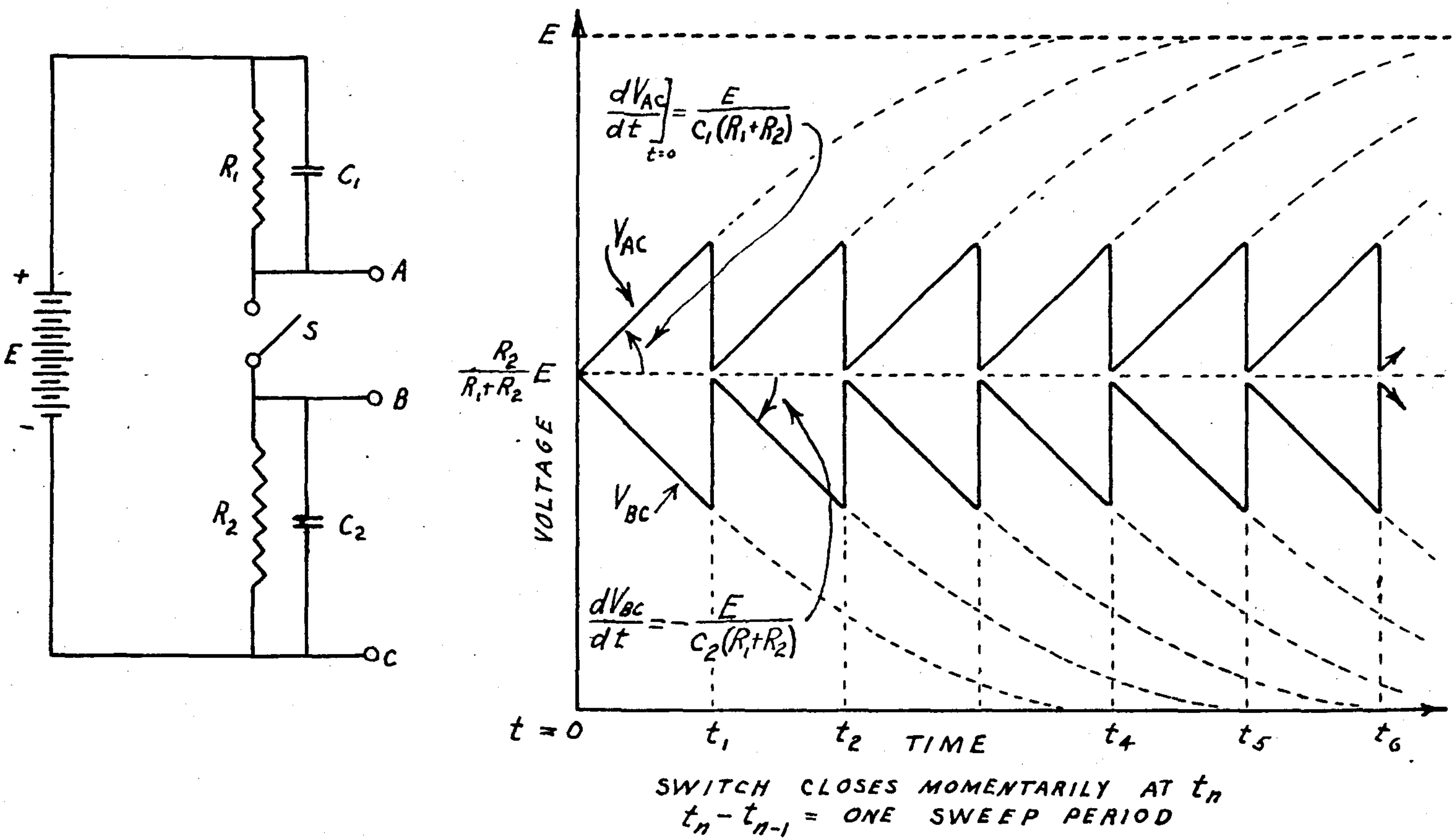
$$(2) \qquad V_{BC} = E \frac{R_2}{R_1 + R_2} \epsilon^{-\frac{t}{R_2 C_2}}$$

$$(3) \qquad V_{AC} = E \left[\frac{R_2 + R_1 (1 - \epsilon^{-\frac{t}{R_1 C_1}})}{R_1 + R_2} \right]$$

Both of these equations represent voltages which are changing linearly with respect to time so long as t is much smaller than RC. Note that the initial rates of change for the two cases are:

$$(4) \qquad \left.\frac{dV_{AC}}{dt}\right]_{t=0} = +\frac{E}{C_1 (R_1 + R_2)} \quad , \quad \left.\frac{dV_{BC}}{dt}\right]_{t=0} = -\frac{E}{C_2 (R_1 + R_2)}$$

Thus if the two condensers are equal in size the variable



Basic Sweep Generator Circuit

Figure 11

portions of the two voltages are balanced with respect to point C regardless of the values of R_1 and R_2.

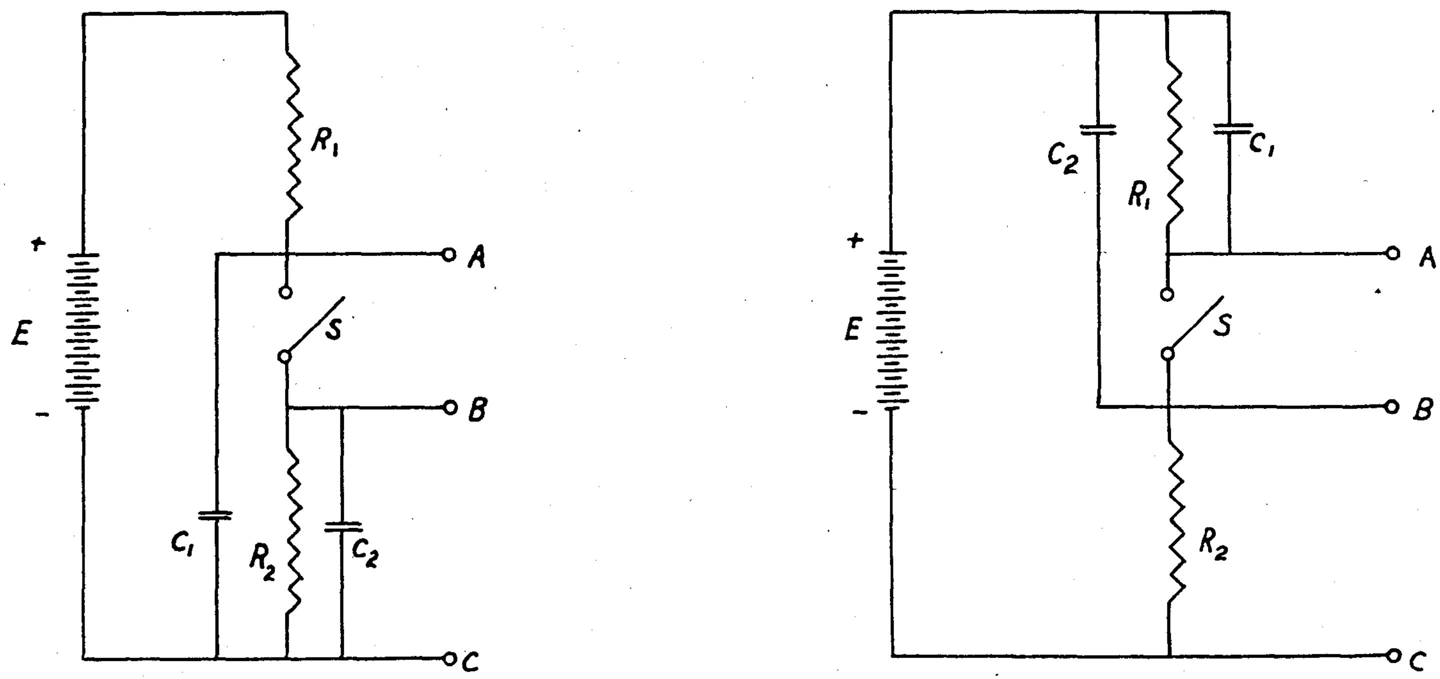
If points A and B are connected to the deflecting plates of a kinescope the total deflection voltage is given by the equation:

$$(5) \qquad V_{AB} = V_{AC} - V_{BC} = \frac{E}{R_1 + R_2}\left[R_1\left(1 - \epsilon^{-\frac{t}{R_1C_1}}\right) + R_2\left(1 - \epsilon^{-\frac{t}{R_2C_2}}\right)\right]$$

Reclosing switch S reduces this voltage V_{AB} to zero, and if the switch is again opened the cycle is repeated to produce a saw tooth voltage.

Examination of equation (5) reveals that the circuit shown in figure eleven is not the only circuit capable of producing the voltage relation set forth in this equation. Figure twelve shows two additional circuits to which equation (5) may be applied. Other circuits of this family can be drawn by changing the fixed voltage points to which the two condensers are returned.

Other circuits can be devised which will produce a sweep voltage which is balanced with respect to ground. Most of them lack two important atributes of the family of circuits to which equation (5) applies. One of these is that the unavoidable capacities from points A and B to ground can be considered as parts of C_1 and C_2 respectively in equation (5). The other factor peculiar to circuits for which equation (5) applies is that these



(a) (b)

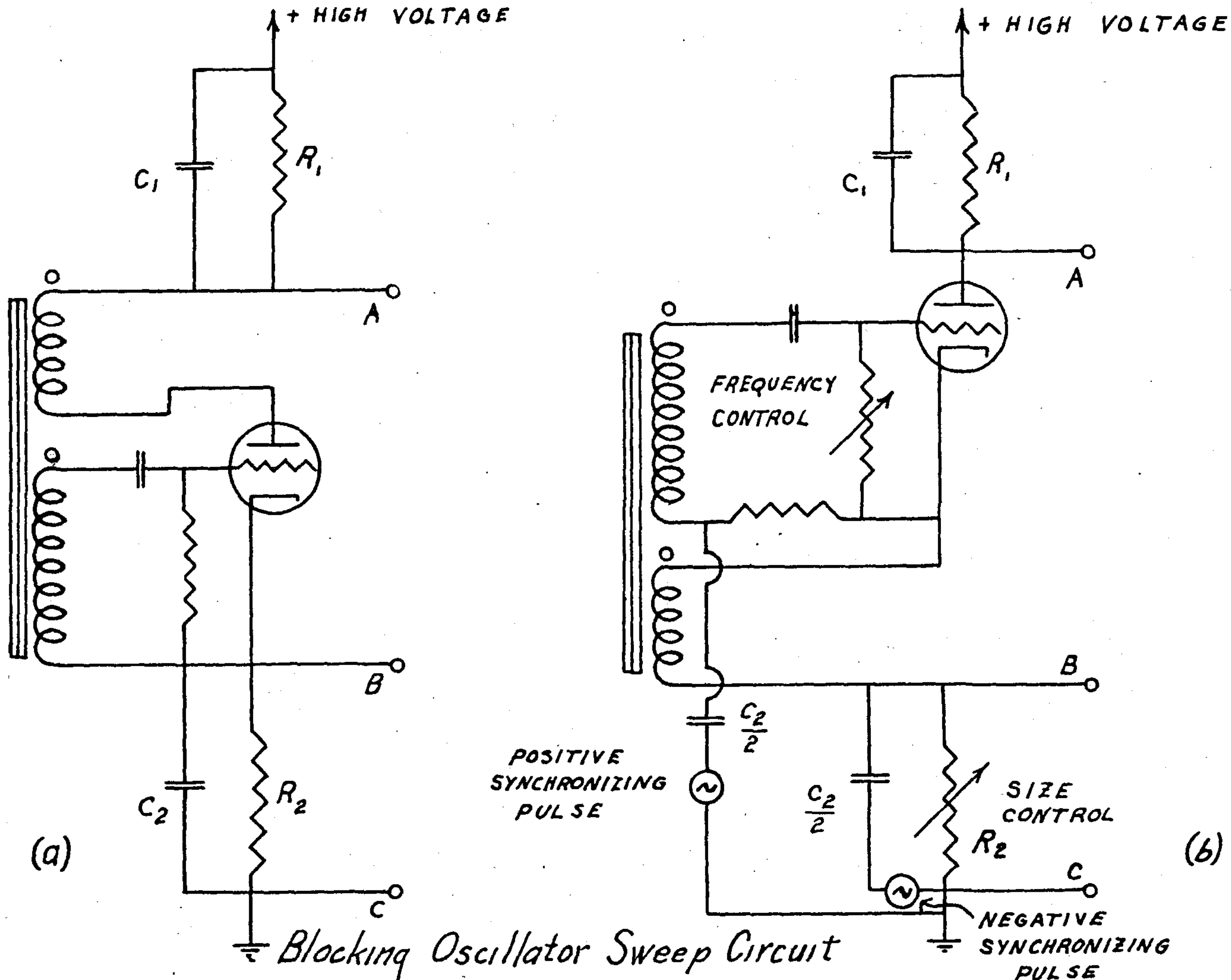
Variations Of Basic Sweep Generator

Figure 12

circuits all have a condenser from each end of switch S to some point of fixed potential. This becomes of great importance if switch S is a vacuum tube the grid of which is to be driven relative to ground.

The switch S in figures eleven and twelve can be either a gas tube or a high vacuum tube. This switch might be arranged so that it is cut off by grid leak bias and turned on by synchronizing pulses. In this arrangement extremely large synchronizing pulses would be required, and the closed resistance of the switch would depend on the amplitude of these pulses. Addition of some type of feedback or regeneration circuit can be used to overcome these two difficulties. Figure thirteen shows a tube connected as a blocking oscillator switch.

The circuit of figure thirteen (a) is the basic circuit of figure eleven with a blocking oscillator inserted as the switch. In thirteen (b) the feedback winding of the transformer has been moved to the cathode circuit to prevent high inter winding voltages in case the switch tube is removed from its socket while the high voltage power supply is on. The grid leak resistor of the switch tube has been made variable to give rough frequency control. Resistor R_2 in the grounded discharge circuit has been made variable to give control of the magnitude of the output voltage. The condenser C_2 has been divided into two parts, and a point for the application of either a



Blocking Oscillator Sweep Circuit
Figure 13.

positive or negative synchronizing pulse has been shown.

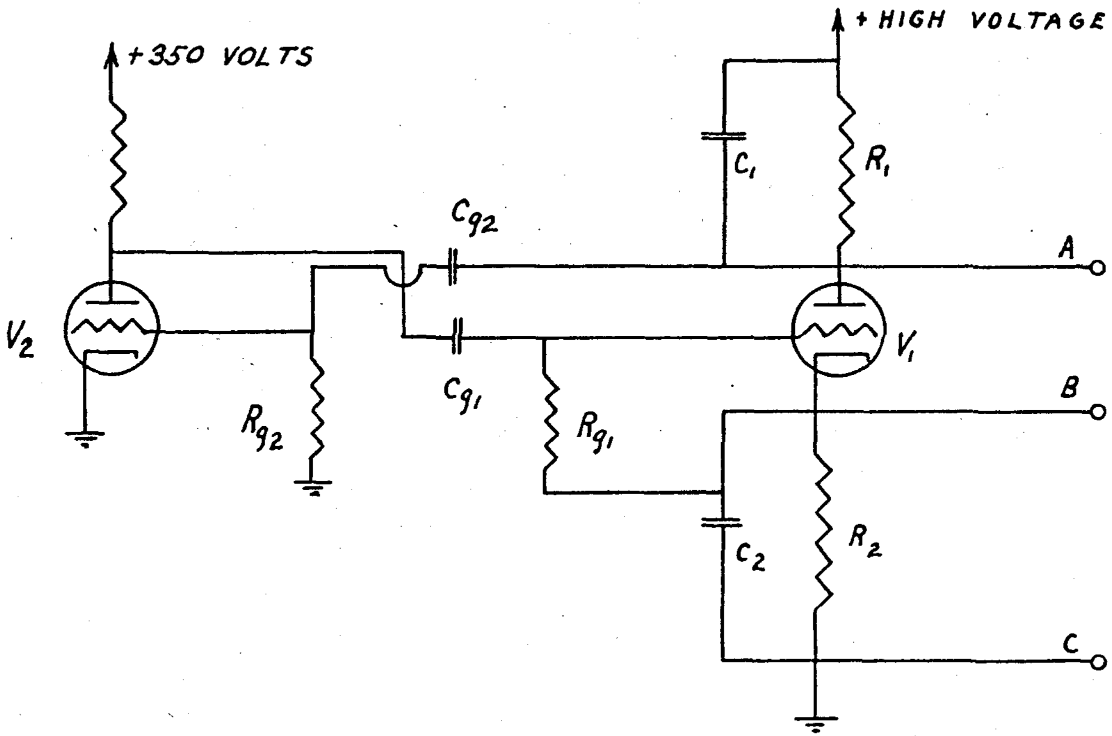
A sweep generator similar to the one shown in figure thirteen (b) was used in an early model of the television receiver produced during this work. It was found possible to use one twin triode of the 6SN7-GT type to produce both horizontal and vertical deflection voltages.

Thus far no mention has been made of the actual power required from the sweep circuit. In general the load into which the sweep circuit works consists of only the two resistors used to supply centering voltages to the deflection plates of the kinescope. These may be made quite large. A value below five megohms is desirable, but in practice these resistors have been made as large as twenty megohms. The use of the circuit of figure thirteen (b) imposes another power requirement. This requirement is that the energy required to charge the tube grid coupling condenser, the energy lost in tube dissipation, and the energy lost in the transformer must be supplied each cycle from the energy stored in the sweep condenser circuit. Since the magnitude of the sweep voltage is fixed by other considerations the amount of energy required at each discharge of the sweep circuit determines the maximum size of the sweep condensers C_1 and C_2. The magnitudes of C_1 and C_2, the sweep period, the required sweep voltage, and the accelerating voltage determine the values for R_1 and R_2. It was found possible to build a satisfactory horizontal sweep generator with a plate input of about two watts. Use of a high quality blocking oscillator transformer and careful selection of the grid condenser and resistor values should make it possible to reduce this to about one watt for a 7JP4 horizontal sweep circuit. Since the power required for the sweep circuit is directly proportional to the sweep frequency the vertical sweep generator power requirements are negligible. In this circuit C_1 and C_2 were selected for ease of component procurement, and the power required was held to about one tenth watt.

Two undesirable features of the circuit shown in figure thirteen (b) are to be noted. First, the frequency control, or hold control as it is called in television circuits, is at a potential much above ground. Since it may be desirable to make this adjustment available from the front panel an insulated mounting bracket and an insulated shaft are required for this control. The second undesirable feature is the requirement of two blocking oscillator transformers for the sweep generator circuits. This last is undesirable only in the sense that the phase inverting function of these transformers may generally be performed by half of a twin triode tube at a lower total cost.

With these factors in mind, the possibility of making the switch S of figure eleven from one half of a multivibrator was investigated. Figure fourteen shows the fundamental multivibrator applied to the switching arrangement of figure eleven. A basic limitation of this circuit is that the peak value of the saw tooth voltage developed between points B and C must be less than the possible change of voltage in the plate circuit of V_2. This is due to the fact that even if C_{g1} is not discharged through R_{g1}point B and the cathode of V_1 will eventually go negative far enough to make V_1 conduct. The B voltage supply of V_2 might possibly be obtained from the accelerating voltage supply, but this is generally undesirable because a current of several milliamperes is necessary in this tube if it is to provide the required steep positive pulse for switching. If the B supply for V_2 is taken from a conventional 300 volt supply a balanced sweep voltage of about 525 volts amplitude can be produced. If an unbalance of about two to one can be tolerated in the two sweep voltages a total sweep voltage of over 700 volts can be produced.

Although the limitation mentioned above makes the multivibrator circuit unuseable in receivers of the a.c.-d.c. type, its very low cost prompted further investigation. In the circuit of figure fourteen the elements R_{g1}, C_{g1}, R_2, and C_2 determine the sweep period. R_{g2} and C_{g2} determine the length of the flyback period.



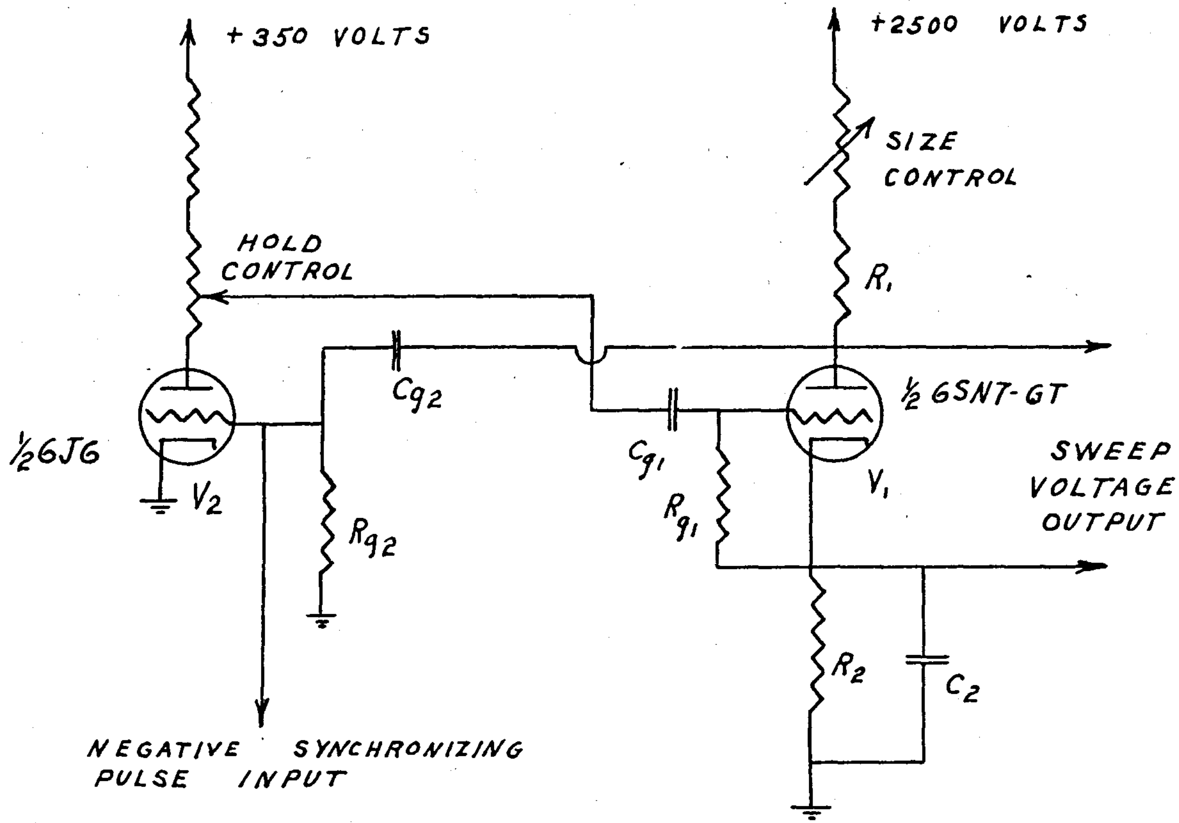
Multivibrator Sweep Circuit

Figure 14

The presence of C_{g2} modifies the required size of discharge capacitor C_1 in the plate circuit of V_1. If R_{g2} is small, as is required for fast fly back, C_{g2} may be increased in size and C_1 may be entirely omitted. The basic circuit then approximates that shown in figure twelve (a).

In figure fifteen a multivibrator circuit is shown in the form in which it was used in one experimental receiver. In this circuit the frequency is controlled by varying the size of the negative pulse used to cut off tube V_1. This arrangement allows the hold control to be at relatively low potential with respect to ground. The size control in this case was placed in the positive discharge resistor. In this circuit, as used, the sweep voltage generated in the cathode circuit of V_1 is smaller than that generated in the plate circuit of this tube when the size control is set for maximum size. Although a change in the size control will require a corresponding change in the setting of the hold control there will not be a change in the voltage balance with change of size. The unbalance of the voltages produced by the circuit of figure fifteen is not great enough to seriously alter the quality of a received picture.

The type of mechanical construction used has an important influence on the costs of the various circuits considered here. Since many of the components in the



Improved Multivibrator Sweep Generator

Figure 15

circuits which have been described are connected so that they operate at a high voltage with respect to ground all of the experimental receivers which were built in this development work had the entire sweep generator built on a sheet of insulating material which could be mounted in a cut-out in the receiver chassis. In later models it was found advantageous to extend the sheet of insulating material for the full length of the chassis and mount the high voltage power supply on it. Since size and position controls for the raster need not be front panel adjustments it is possible to mount them on this insulated panel such that they are available from the rear of the set with an insulated screw driver.

The multivibrator of figure fifteen is synchronized by a negative pulse applied to the grid of tube V_2. During the sweep period the grid of V_2 is slightly positive with respect to its cathode, and V_2 is conducting heavily. Because of this condition the negative pulse on the grid of V_2 is amplified by a factor of about thirty before application to the grid of V_1. Due to the amplifying action of V_2 it is possible to use one less synchronizing signal amplifier than would be necessary with a blocking oscillator type sweep generator.

With the type of sweep generator which has been considered here the cathode of the sweep generator tube operates at an average potential 1000 to 2000 volts above ground. A separate winding on the power transformer is used to heat the filament for this cathode. Where two sweeps are to be generated in one twin triode the insulation between filament and cathode is of great importance. The peak voltage between cathodes is equal to the peak sweep voltage generated in either cathode circuit. The filament is bypassed to ground by a condenser large enough to hold it at some average level determined by the leakage resistance between the filament and the two cathodes and that between the filament and ground. According to tube ratings the filament cathode voltage should not be allowed to exceed about 90 volts on a 6SN7-GT tube. A balanced sweep of 360 volts peak to peak can be produced by each triode section of this tube without exceeding the filament cathode rating. It has been found in practice that since the series resistances in the circuit are high and since the energy storage capacity of the circuit can be kept low enough to prevent destructive arc over it is possible to operate the 6SN7-GT with peak heater cathode voltages of 150 to 200 volts. The strain on this insulation can be greatly reduced if the filament voltage is very slightly reduced and if the high voltage power supply is arranged to use a slow heating diode which will not conduct until the sweep tube is at operating temperature.

It is not possible to give any data on the peak value of the sweep voltages which can be generated with a twin triode such as the 6SN7-GT as this type of tube is generally not rated for the high voltage low current type of operation encountered in this sweep circuit. A large number of 6SN7-GT tubes manufactured by different companies have been tried in the sweep circuits described here. Only one tube has failed in service. This tube, which had previously been used in an amplifier for a number of months, failed due to an internal short formed around a flake of cathode coating.

If a sweep voltage of over six hundred volts peak is required it is probable that best results could be obtained by use of two separate tubes with two separate heater windings on the power transformer to supply their filaments. If this sweep circuit is to be used for a receiver to be operated from 115 volt direct current sources the accelerating potential supply may consist of a radio frequency high voltage generator and a rectifier. The filament of the rectifier will normally be supplied by a winding on the radio frequency transformer. In this case two small battery type triodes may be used for sweep generator tubes, and their filaments can be supplied by windings on the radio frequency transformer.

THE HIGH VOLTAGE POWER SUPPLY

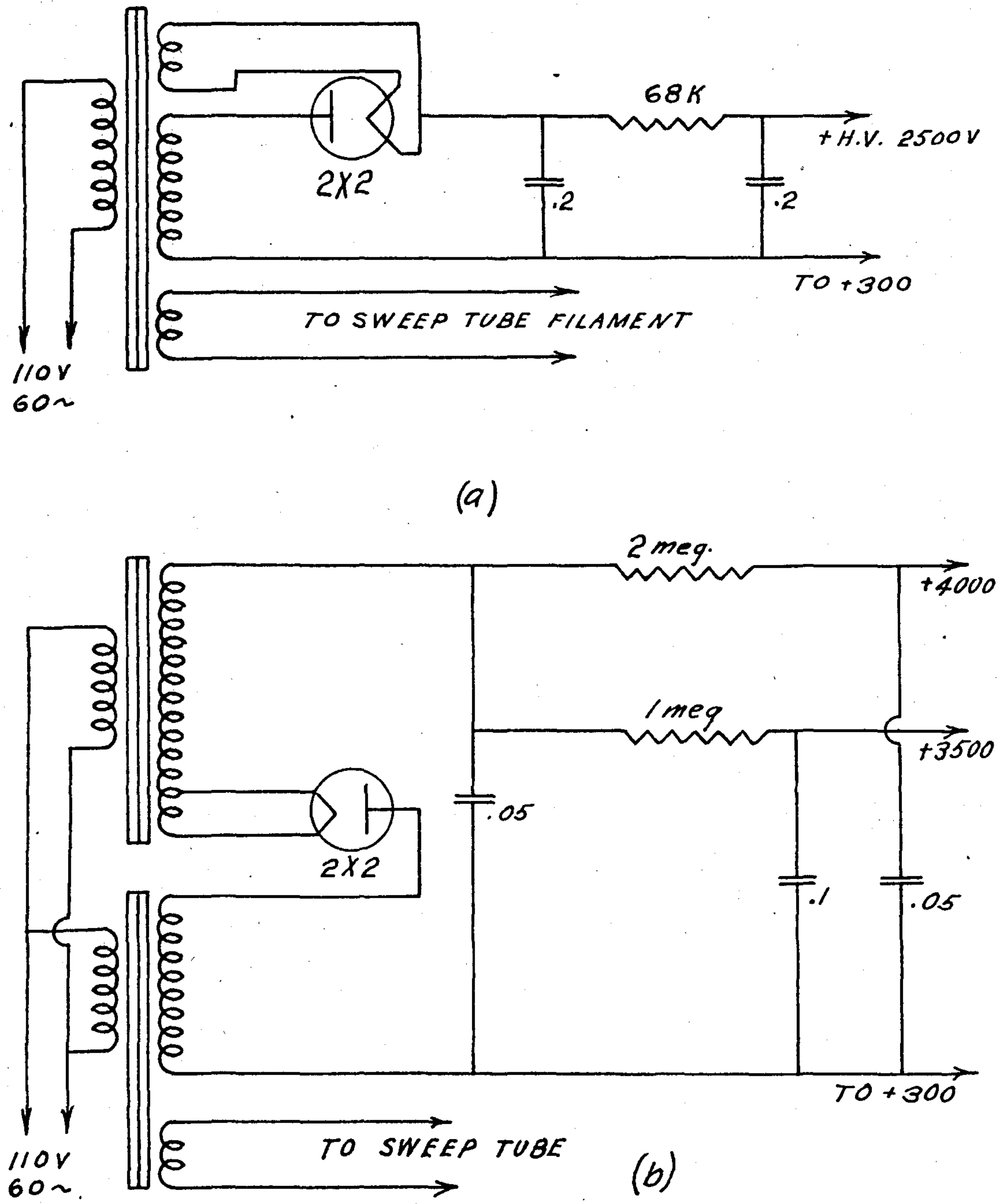
Since the high voltage power supply requirements of a small television set are not hard to meet the design of the high voltage power supply requires no trick circuitry. It is only necessary to put down all the available types of power supplies which will meet the current and voltage requirements of the set, compute their different costs, and select the least expensive supply. The various current drains on the supply are approximately as follows:

Kinescope	100-300 Microamperes
Bleeder	200 "
Sweep Generator	700-1000 "
Maximum total	1.5 milliamperes

The voltage required depends on the contrast range and the fineness of focus desired. For the 7JP4 tube a high voltage supply furnishing 4000 volts to the tube second anode is adequate to secure a good picture. Many television sets now available use the 7JP4 tube with a lower accelerating voltage with a resultant loss of picture quality. A high voltage supply furnishing much more than 4000 volts requires condensers with voltage rating in excess of 5000 volts, and this results in a considerable increase in cost.

If the receiver is to be operated from a source of direct current the high voltage power supply will be of the radio (or audio) frequency generator type, but if a source of alternating current is available the cost factor favors a conventional power supply with an iron core transformer. A worth while gain in output voltage of the supply can be obtained by returning the negative terminal of the high voltage supply to the positive terminal of the regular B voltage supply of the receiver. If the gain in output voltage is not needed this connection may be used to make available another 300 volts which may be used in the resistor condenser filter to secure better filtering.

Since the current drain on the high voltage power supply is very low the use of an inductor in the filter circuit is impractical. Sufficient filtering could be obtained with one large condenser, but the same amount of filtering can be obtained at lower total cost by the use of two smaller condensers and a resistor in a pi type filter. The filtering action of this circuit is approximately proportional to the product of the output condenser and the series resistor. For this reason it is well to consider the possibility of making the high voltage winding of the power transformer considerably larger than necessary to get the desired output voltage. The series resistor of the filter can then be increased, and the same filtering action can be obtained with smaller condensers. One power supply which was used in experimental receivers is shown is figure sixteen (a). The voltage output of this receiver was not as high as desired. Some sample multiple condensers procured for test purposes and the addition of another small transformer made possible the design of a much more satisfactory power supply shown in figure sixteen (b). The condenser here is a .05/.05/.1 microfarad unit rated at 5000 volts. The 4000 volt supply is connected to the kinescope second anode and to the bleeder stick from which positioning and focusing voltages are taken. The 3500 volt supply is used to generate the required 700 volts of sweep signal. A small resistor is placed in the rectifier filament lead. This assures that all other tubes in the set will be hot before the high voltage is applied. The initial surge voltage on condensers in the circuit is thus reduced, and burning of the screen by a bright stationary spot is prevented.



High Voltage Power Supply

Figure 16

THE LOW VOLTAGE POWER SUPPLY

Unless the television receiver is to be built to operate a.c.-d.c., the lowest cost can be achieved by use of a transformer for the low voltage power supply. This is not apparent unless the cost of the required filament bypass condensers is counted as a part of the power supply cost. With a transformer supply the filaments of the tubes can all be operated with one end grounded. (The sweep tube is, of course, an exception.) With one end of each tube filament grounded there is little chance of radio frequency feed back in the filament circuit. This is not the case where a number of filaments are operated in series.

Because of the low total number of tubes required in the circuits used in this development it was possible to operate the receiver with a B supply current drain of about 80 milliamperes at 300 volts.

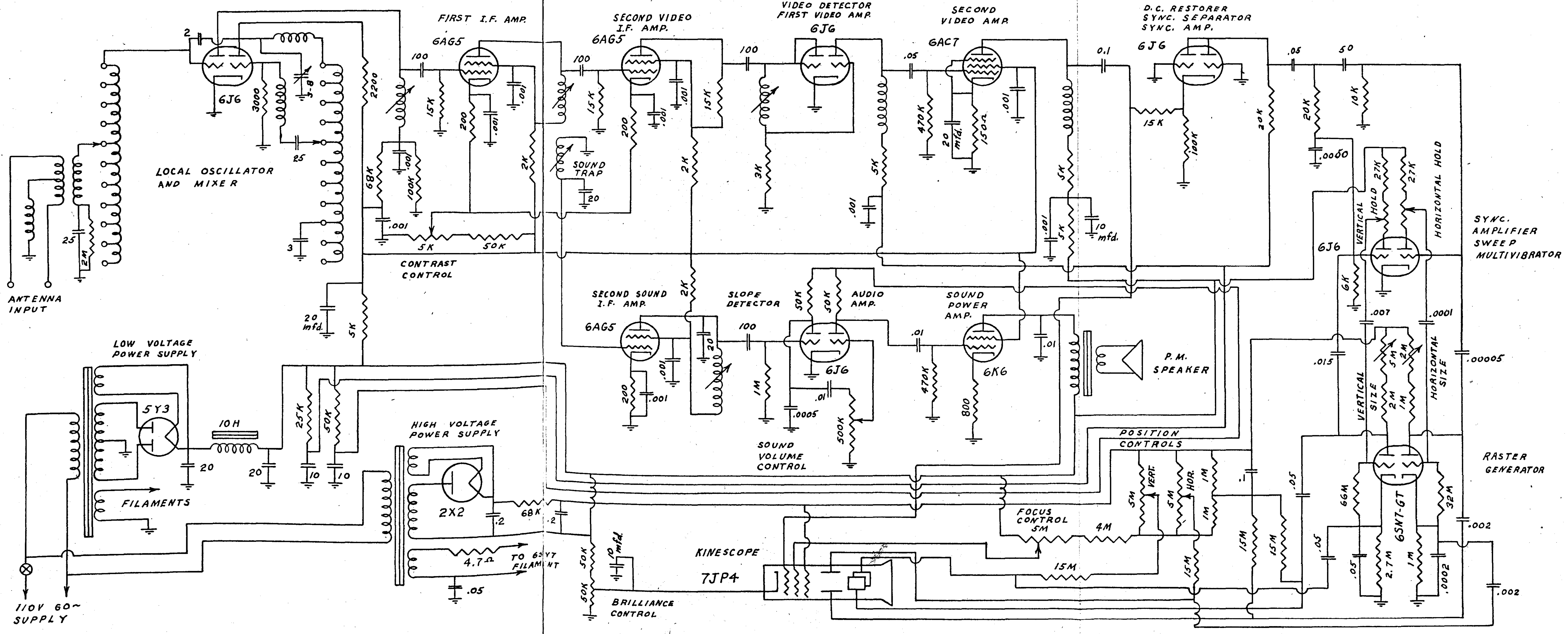
By mounting the transformers and chokes of the power supply beneath the front part of the chassis the effect of their stray field on the kinescope can be greatly reduced. In one early model of the receivers built in this investigation these units were mounted in more conventional positions at the rear and above the chassis. In this case a rather extensive magnetic shield around the neck of the kinescope was required to prevent defocusing and distortion of the raster.

POSSIBLE MODIFICATIONS

In figure seventeen a diagram of one completed model of the receiver is shown. This receiver is capable of giving good results on signals from local transmitters, and if used with a good antenna it will produce useable pictures at a distance of 35 to 40 miles from the transmitter over relatively flat country. Tests indicate that it is probably not the least costly device which will produce a picture from a local transmitter. Nor will it produce as good a picture as is desired where signal strengths are low. For these reasons it is well to investigate what may be done to lower cost for a receiver for strong signals and what may be done to improve performance where signals are weak.

Little experimental work has been done on a less costly receiver for shorter range. Since in the design of the receiver described here the costs of the circuits for the individual functions were reduced as much as possible the natural line of approach is to investigate the possibility of leaving out some of the functions performed in this receiver. It is possible that the receiver cost could be reduced by changing the basic receiver design from a superhetrodyne to a tuned radio frequency type. The next logical step in the reduction of the basic receiver is a simple crystal detector circuit.

Complete Developmental Receiver Using Low Cost Circuitry
Figure 17



In this circuit the antenna signal would be coupled to a broadly tuned circuit resonant at the frequency of the desired channel. A fixed crystal detector would rectify the voltage across this tuned circuit and the resulting signal would be coupled to a high gain video amplifier. A 4.5 megacycle trap in the video output would serve to remove the sound signal from the video, and slope detection of the voltage across the trap would provide the required sound signal. Sweep circuits might be those used in the receiver described in this paper.

There are a large number of improvements which can be made in the receiver described in this paper. For each improvement there is a corresponding penalty in the form of a cost increase. This fact makes it especially desirable to develope those improvements which can be added to the basic receiver when the receiver is to be used in an area where television reception is poor.

The sensitivity of the receiver could be increased by addition of amplifier stages in both video and sound channels. A more desirable method of increasing the sensitivity is through the addition of a tuned radio frequency amplifier stage ahead of the receiver, for this also provides some much needed additional image and intermediate frequency rejection. To provide this increase in sensitivity in such form that it could be purchased for areas of low signal strength, or omitted in high signal strength areas a self contained signal booster has been developed. The booster is essentially a tuned radio frequency amplifier with a channel switch which varies the tuning for the various channels. An a.c.-d.c. power supply is built into the booster.

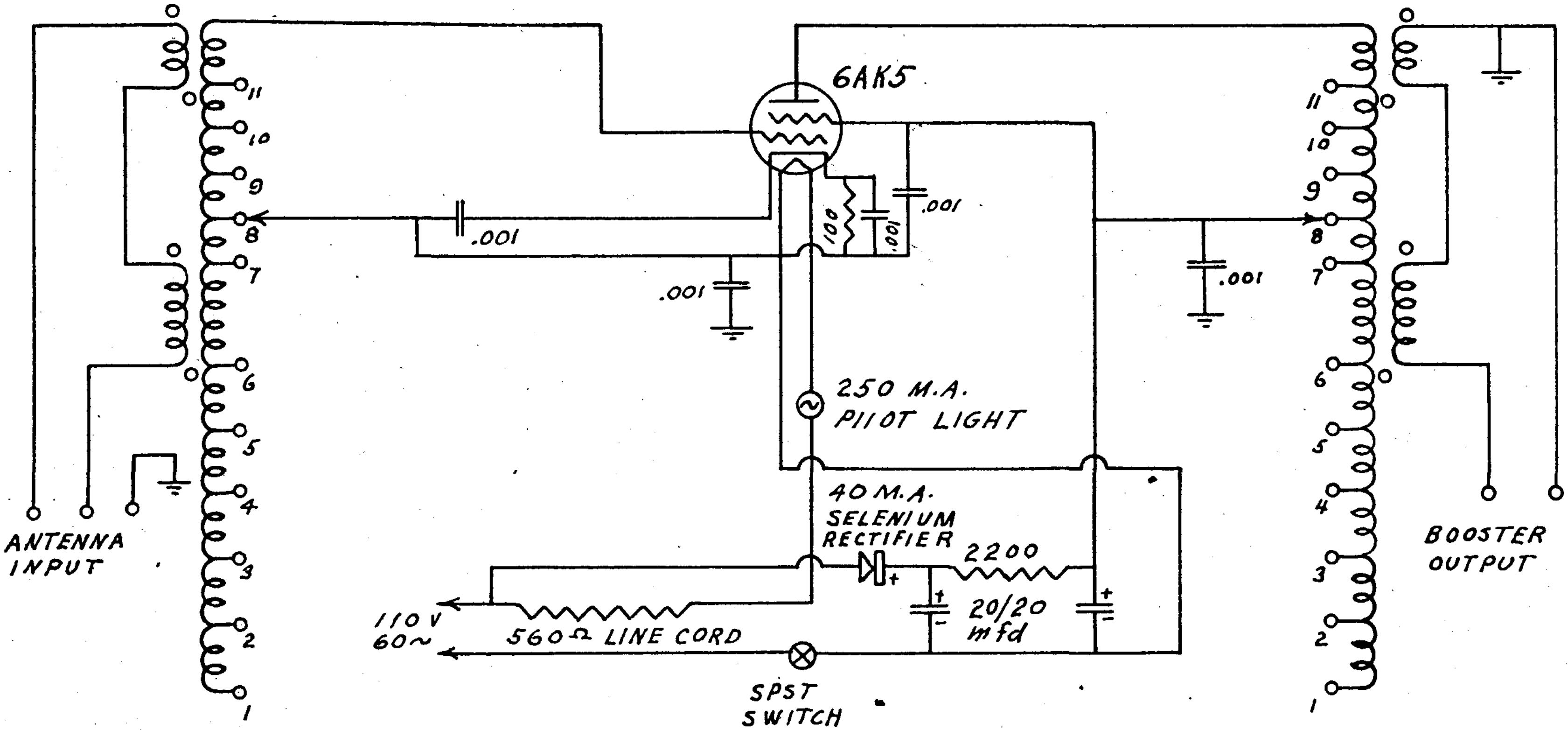
In the design of a booster the problem of coupling the antenna energy into the tube grid circuit is encountered just as it was in the development of the frequency converter. In the output circuit of the booster the inverse problem, that of coupling energy from the plate circuit to a low impedance line, is met. The best low cost tube available for a high frequency booster is probably the 6AK5. If noise were the limiting factor consideration should be given to amplifiers such as the cathode coupled wide band amplifier.[3] In this case, however, the main consideration is a high gain bandwidth product, and for this the tube used must have high transconductance, low input and output capacities, and no more input loading than necessary to give the required bandwidth at the highest frequency used. Experimentally determined values of input loading for pentodes follow the law $R_i - 1/kf^2$ where R_i is the equivalent shunt resistance presented by the tube.[1] For the 6AK5 a value of 1.2×10^{-12} may be used for k in the television frequencies. Values of k for other tubes in this frequency range vary from 2.5×10^{-12} for the 6AG5 to 20×10^{-12} for the 6AC7.

In order to secure the low input loading of the 6AK5 advantage must be taken of the double cathode lead provided to reduce the effect of cathode lead inductance.

Figure eighteen is the circuit diagram of the completed booster. It will be noted that a compromise coupling scheme has been used in both input and output circuit to improve the gain in the low frequency channels. The low channel coupling coils wound on inductors L_6 and L_{17} have high distributed capacity such that they present small reactance in the high frequency channels. In the high frequency channels the coupling is inductive in inductors L_1 and L_{12}. Input loading of the order of 1000 to 2000 ohms in the grid circuit and close coupling to a 300 ohm output line in the plate circuit make it unnecessary to load these circuits with additional resistance to secure the required band width in the high channels.

In the low channels the output coupling is obtained by mutual inductance between primary and secondary in L_{12} and L_{17}. It is necessary to ground one side of the output line to prevent undesired loading of the output circuit by the radiation resistance of the two output wires acting in parallel as a long wire antenna capacitively coupled to the plate of the tube.

Input coupling on the low channels may be inductive through L_1 and L_6 if a signal voltage which is balanced with respect to ground is available. If an unbalanced



Signal Booster
Figure 18

signal is available as in a coax cable the signal may be coupled in through the stray capacity between primary and secondary in L_1 or L_6.

In the low frequency channels the loading in plate and grid circuits is not great enough to give the required bandwidth. For this reason the input and output circuits are stagger tuned. Greatest stability is obtained with the grid circuit tuned to a frequency lower than the resonant frequency of the plate circuit.

With the booster shown here it is possible to get a voltage gain of ten on all television channels. In the high frequency channels the bandwidth is greater than necessary. In the low frequency channels the bandwidth can be made greater than is necessary for the receiver which has been described in this paper.

Several other optional additions to the basic receiver are possible. By addition of a double diode and a substitution in transformer types it is possible to convert the sound detector to a frequency discriminator type. In this case the original slope detector may be used as an additional stage of high gain audio amplification. Improved synchronizing of the trace in the raster and a resulting improvement in definition can be obtained by addition of a twin triode as two cascade stages of synchronizing signal amplifier. If this is done the synchronizing signal separator may be designed to clip at a very low level, and thus the possibility of synchronization on noise is reduced.

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